AUGUST · 1953

Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

FOCUSING ULTRASONICS TRANSDUCER

Brush Electronics Company

This impressive wave pattern portrayed by thousands of bubbles being produced by the tremendous energy radiated from a barium nate ceramic transducer submerged in water.

Volume 41 Number 8

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The Institute of Radio Engineers

NEW "M" TYPE TOROIDS Maximum Q Minimum Size

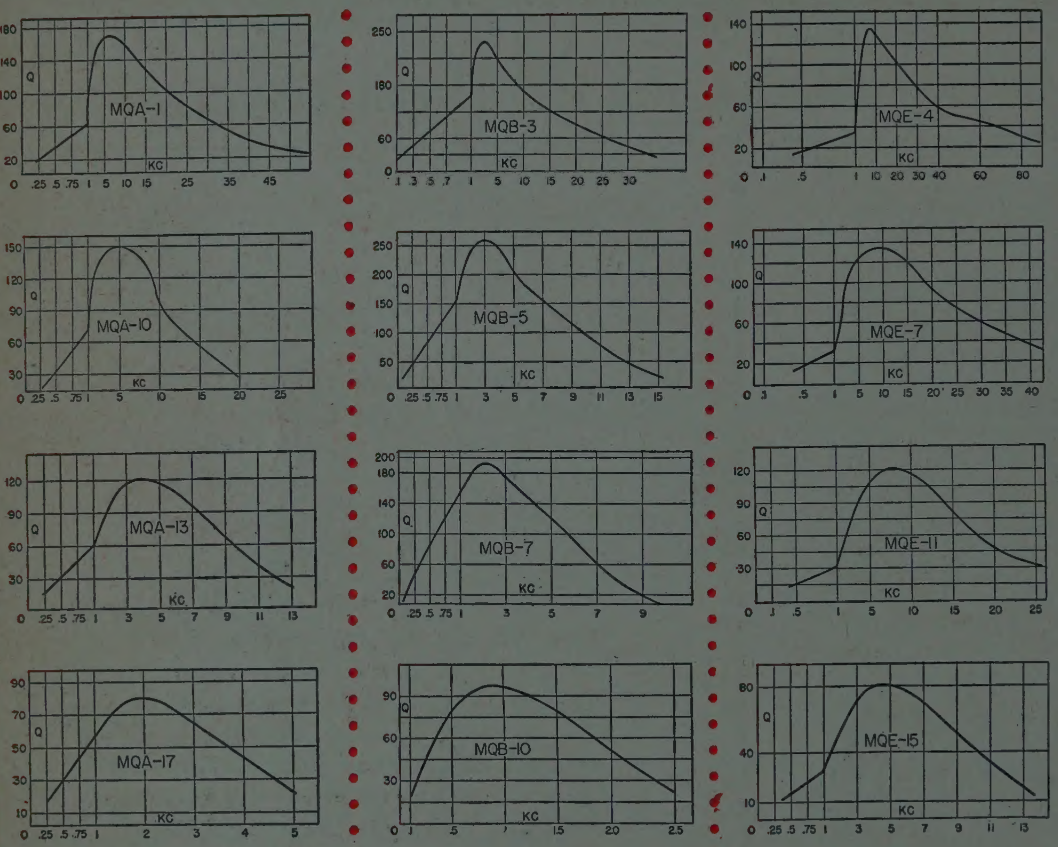
UTC Permalloy Dust Toroids have been the standard of the industry for over 15 years. The **MQ series** of coils provide the highest Q factor in their class (see curves below), with miniaturized dimensions. All units are **hermetically sealed** to MIL-T-27 Specifications.

The **stability** is excellent. For the MQE-7 the inductance change is less than 1% for voltages from .1 to 3 volts. The MQA-13 change is less than 1% for applied voltages from .1 to 20 volts. The MQB-5 change is less than 1% for applied voltages from .1 to 50 volts. DC is permissible through the coil (values listed below). Inductance is virtually independent of frequency temperature and vibration.

Hum pickup is extremely low due to the toroidal winding structure, with windings uniformly spread over the core. The case is of high permeability, affording additional shielding such that close spacing of units can be effected, the coupling attenuation being approximately 80 DB.

Other values of inductance than those listed are available on special order at the price of the next higher listed value.

TYPICAL Q CURVES



MQA TYPES

Type No.	Inductance	*DC Max.
MQA-1	7 mhy.	250
MQA-2	12 mhy.	200
MQA-3	20 mhy.	150
MQA-4	30 mhy.	125
MQA-5	50 mhy.	100
MQA-6	70 mhy.	80
MQA-7	120 mhy.	60
MQA-8	.2 hy.	50
MQA-9	.3 hy.	40
MQA-10	.5 hy.	30
MQA-11	.7 hy.	25
MQA-12	1 hy.	20
MQA-13	1.5 hy.	17
MQA-14	2.5 hy.	13
MQA-15	4 hy.	10
MQA-16	6 hy.	9
MQA-17	10 hy.	7
MQA-18	15 hy.	5
MQA-19	22 hy.	4

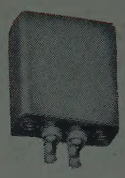
*This value of D.C. (MA) will drop the coil inductance 5%. Values of D.C. below this will show proportionately (linear) less inductance drop. For example, MQE-1 will drop ½% in L with 13.5 MA.

MQB TYPES

Type No.	Inductance	*DC Max.
MQB-1	10 mhy.	400
MQB-2	30 mhy.	250
MQB-3	70 mhy.	170
MQB-4	120 mhy.	120
MQB-5	.5 hy.	60
MQB-6	1 hy.	40
MQB-7	2 hy.	30
MQB-8	3.5 hy.	22
MQB-9	7.5 hy.	16
MQB-10	12 hy.	11
MQB-11	18 hy.	9
MQB-12	25 hy.	8

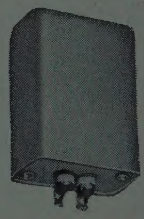
MQE TYPES

Type No.	Inductance	*DC Max.
MQE-1	7 mhy.	135
MQE-2	12 mhy.	100
MQE-3	20 mhy.	80
MQE-4	30 mhy.	65
MQE-5	50 mhy.	50
MQE-6	70 mhy.	40
MQE-7	100 mhy.	35
MQE-8	150 mhy.	30
MQE-9	.25 hy.	22
MQE-10	.4 hy.	17
MQE-11	.6 hy.	14
MQE-12	.9 hy.	12
MQE-13	1.5 hy.	9
MQE-14	2 hy.	8
MQE-15	2.8 hy.	7.2



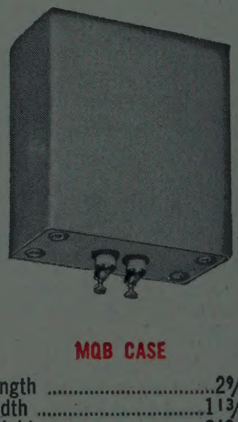
MQE CASE

Length	$1\frac{1}{16}$"
Width	$\frac{1}{2}$"
Height	$1\frac{17}{32}$"
Unit Weight	1.5 oz.



MQA CASE

Length	$1\frac{9}{32}$"
Width	$1\frac{1}{16}$"
Height	$1\frac{23}{32}$"
Unit Weight	4 oz.



MQB CASE

Length	$2\frac{9}{16}$"
Width	$1\frac{13}{16}$"
Height	$2\frac{13}{16}$"
Unit Weight	14 oz.

United Transformer Co.

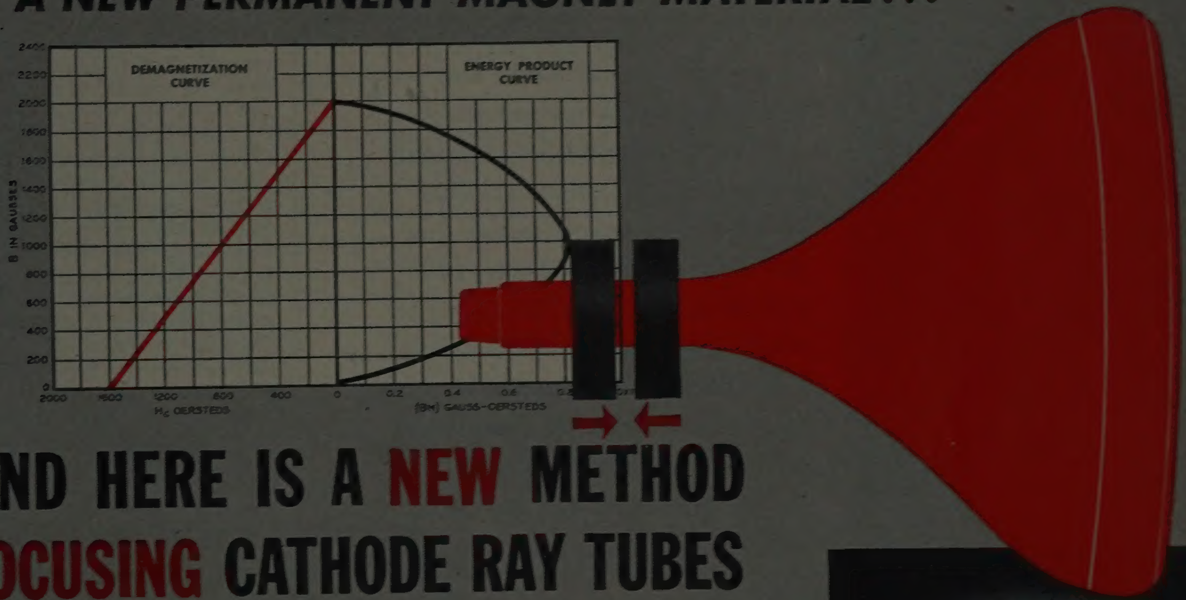
150 VARICK STREET NEW YORK 13, N. Y.

EXPORT DIVISION: 13 EAST 40th STREET, NEW YORK 16, N. Y., CABLES: "ARLAB"

NEW!

MAGNADUR

A NEW PERMANENT MAGNET MATERIAL . . .



. . . AND HERE IS A NEW METHOD OF FOCUSING CATHODE RAY TUBES

Here is another Ferroxcube "first": a permanent magnet material with outstanding magnetic characteristics—*and no critical materials are involved in its manufacture.* Magnadur's extremely high coercive force and unusually high resistance to demagnetization permit entirely new magnet designs.

Magnadur will be produced in a variety of shapes. Production for the current year is concentrated on Magnadur toroids—developed specifically for TV focusing ring magnets.

Magnadur focus rings provide a real answer to TV focus problems. The double lens system, which is focussed by adjusting the relative position of two toroidal magnets, reduces stray fields to a minimum and provides a highly symmetrical field. Maximum sharpness and spot symmetry are assured.

Technical information will be sent upon request. Ferroxcube engineers are at your service for consultation. We'll be pleased to have you call or write.

SEND FOR BULLETIN

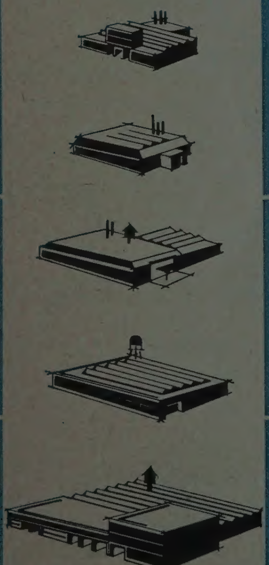
EXAMPLE OF
A COMPACT
TV FOCUSING DEVICE
USING
MAGNADUR
RINGS

FERROXCUBE CORPORATION OF AMERICA
• *A Joint Affiliate of Sprague Electric Co. and Philips Industries, Managed by Sprague* •
235 EAST BRIDGE STREET • SAUGERTIES, NEW YORK

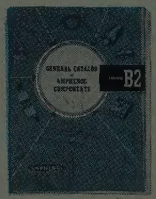
AMPHENOL

builds to the future of electronics

Quality components, new ideas—these are the contributions Amphenol has made to the electronics industry for over 20 years. During that time Amphenol's production of components has gradually increased until today Amphenol represents the largest single source for users of AN connectors, RF connectors, radio sockets and coaxial cables. And in the wonderful future that lies before the entire electronics industry, Amphenol is proud to claim a part. The record of past accomplishments, the ingenuity of the Amphenol engineering staff, plus the seemingly unlimited applications possible for new components insure that Amphenol will always provide the necessary parts for new electronic equipment. "Building to the future of electronics" is more than a slogan at Amphenol—it expresses our belief that before the electronics industry lies a golden age.



Amphenol has also steadily expanded their production facilities through the years: there are now five modern plants to better serve you with quality electronic components.



All products illustrated on the opposite page are listed in the General Catalog, B-2. If more complete product data is required, the inside cover page, B-2, will provide listings of the special catalogs and bulletins.

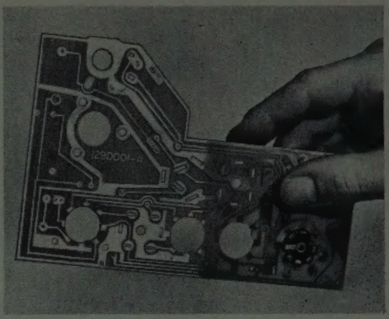
AMPHENOL

AMERICAN PHENOLIC CORPORATION • *chicago 50, illinois*

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

Activated Rosin Flux



Insulation characteristics comparable to polyethylene have been successfully combined with extremely fast "take" in a new rosin flux by the **London Chemical Co., Inc.**, Dept. P17, 325 W. 32nd St., Chicago 16, Ill. According to the announcement, this flux is devoid of any free acid and is completely non-corrosive. These features meet two particularly important needs in present electronics applications: (1) fast fluxing action and smooth, even coating necessary in the dip tinning of printed circuits; (2) elimination of breakdown due to corrosion in the soldering of uhf components. Additionally, this new rosin flux is unusually resistant to high temperatures. Thus, its ability to withstand prolonged application of heat enables it to solder through badly oxidized copper surfaces and to solder hard-to-bond metals and alloys such as cadmium, nickel, and nichrome steel.

The 1953 Audio Fair

According to the announcement last week by Harry N. Reizes, Fair Manager, it is expected that more than 20,000 music lovers, audiophiles, and sound engineers will visit the Hotel New Yorker, October 14, 15, 16, and 17, to make the 1953 Audio Fair the great public exhibit in audio Hi-Fi history. Using the attendance at previous Fairs in New York, Chicago, and Los Angeles as a basis of comparison, this estimate of attendance is very conservative.

Paralleling a new high in attendance will be a record-breaking number of exhibitors, since more than one hundred have already engaged space. Included are a number of major manufacturers, among them RCA, Hallicrafters, I.D.E.A., National Company and others who have chosen the Audio Fair as an effective means of introducing publicly the Hi-Fi equipment they are manufacturing this year for the first time.

Indicative of industry growth and the acceptance of the New York Audio Fair as the official meeting place of manufacturers, dealers, and hobbyists, is the fact that the Fair this year will occupy three floors of the New Yorker in comparison to the two

(Continued on page 21A)

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 20A)

floors which have sufficed in previous years.

Reflecting the penetration of high quality reproduced sound and music into all aspects of home and industry use, the 1953 Fair will be built around the theme of "Audiorama."

"Audio has become a distinct entity in the panorama of our daily existence," Mr. Reizes said in his announcement. "We feel that the word 'Audiorama' dramatizes the extent to which audio has earned acceptance as one of the prime amenities in the American scene. Hence its choice as the theme of the 1953 Audio Fair," he concluded.

Joining with Mr. Reizes in announcing completion of plans for the country's foremost audio event were officials of the Audio Engineering Society, whose annual convention is scheduled to coincide with the Audio Fair. Mr. Jerry Minter, convention chairman, stated that this year the Society will entertain a greater array of technical papers on audio subjects than has ever before been presented in a single meeting.

In keeping with the policy established with the first Audio Fair five years ago, the 1953 event will be open to all those interested, both professional and amateur, free of charge.

New TV Studio Equipment

Film editing equipment, a dual-disc slide projector, and a large screen projection TV system, three new aids designed to better studio programming for television stations, now are available from the **Engineering Products Dept., RCA Victor Div., Radio Corp. of America,** Camden, N. J.

The film editing equipment consists of film splicers, rewinds, viewers, measuring machine, editing tables, film or slide storage cabinets, reels, cans, and film polishers and cleaner.

The new dual-disc slide projector (RCA Type TP-3A) is for either glass or cardboard mounted 2-inch square slides. Push-button controls permit remote operation from the audio/video console. The projector has a 150-watt lamp, two five-inch lenses, and two slide turrets which are operated by individual motors. It can be mounted on the conventional RCA multiplexer stand if desired.

The large-screen television projector (RCA Type PT-100), has overcome the problem of obstructing the view of the studio audience and enables the viewer to follow the program as rapid switching takes place between sets and scenes. It also gives a producer a check on the technical quality of the pickup. Descriptive literature may be secured by writing to the **Broadcast Equipment Section, Radio Corp. of America,** Camden 2, N. J.

(Continued on page 26A)

AN CONNECTORS
AMPHENOL is the leading manufacturer of approved AN connectors. These feature premium material, and careful inspection assures that each connector measures up to and beyond specifications.

RF CONNECTORS
AMPHENOL RF connectors provide never failing continuity and extremely low-loss—are unsurpassed for mechanical design and electrical efficiency.

POWER CONNECTORS
AMPHENOL'S lightweight and compact power connectors are built to stand rough handling. They are 100% shockproof and can be relied upon for electrical connections in all appliances and machinery.

AUDIO CONNECTORS
AMPHENOL microphone connectors have been standard with leading microphone manufacturers for years. Providing unique interchanging coupling rings, they give mating connections at every junction.

RACK & PANEL CONNECTORS
AMPHENOL 26 series rack and panel connectors have the added strength needed for their efficient operation and safety features which include interlocking barriers to prevent accidental shorting.

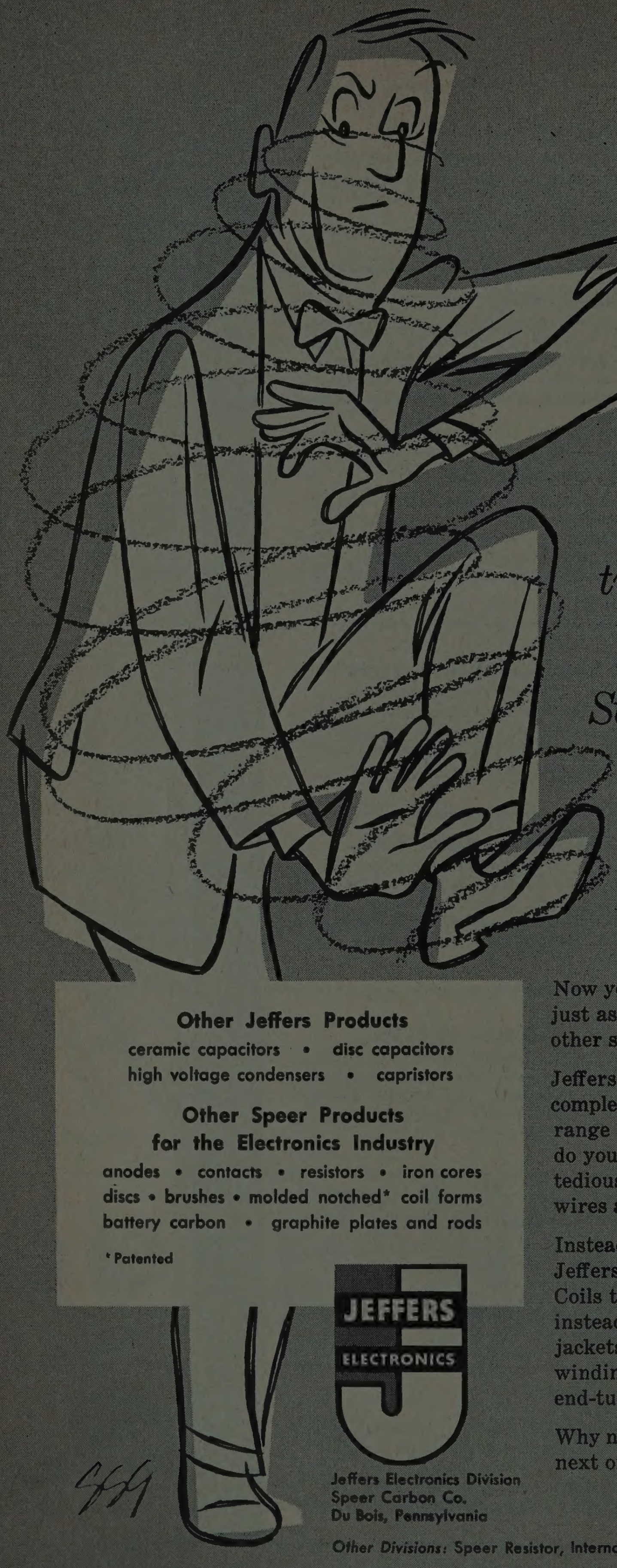
BLUE RIBBON CONNECTORS
AMPHENOL Blue Ribbon connectors represent a new solution to the problem of providing quick disconnection for electronic sub-assemblies, incorporate gold finished contacts and new sturdy dielectric.

RG TYPE COAXIAL CABLES
AMPHENOL RG coaxial cables are made with low-loss polyethylene dielectrics. Precision extrusion guarantees strict end-to-end uniformity—constant inspection insures top quality.

MINIATURE CONNECTORS
AMPHENOL miniature connectors provide high-quality interconnection of miniature electronic equipment. They are made with the same skill and care that characterizes all AMPHENOL components.

SOCKETS
AMPHENOL MIP sockets have unique construction features. The plate is molded directly into the bakelite body, cannot come loose or vibrate—speeds up production, reduces breakage.

AMERICAN PHENOLIC CORPORATION • chicago 50, illinois
AMPHENOL



All wound up

trying to make your own R. F. Choke Coils?

Standardize on **Jeffers R.F. Choke Coils-**

save time, labor and expense

Now you can stock a wide range of R. F. choke coils just as you do resistors, capacitors and other similar components.

Jeffers Electronics is ready to deliver to you a complete line of R. F. choke coils with the widest range of inductance values available. No longer do you have to waste time, labor and money on slow, tedious hand assembly from miscellaneous forms, wires and coatings.

Instead you receive a standardized product from Jeffers, completely assembled and ready for use. Coils that are well made, too. Insulated copper wire instead of bare wire for windings . . . husky, molded jackets instead of those fastened by glue. All windings are soldered to leads . . . shorted end-turns completely eliminated.

Why not give Jeffers R. F. choke coils a try on your next order? *Write today for our specification sheets.*

Other Jeffers Products

ceramic capacitors • disc capacitors
high voltage condensers • capristors

Other Speer Products for the Electronics Industry

anodes • contacts • resistors • iron cores
discs • brushes • molded notched* coil forms
battery carbon • graphite plates and rods

* Patented

JEFFERS ELECTRONICS

Jeffers Electronics Division
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Du Bois, Pennsylvania

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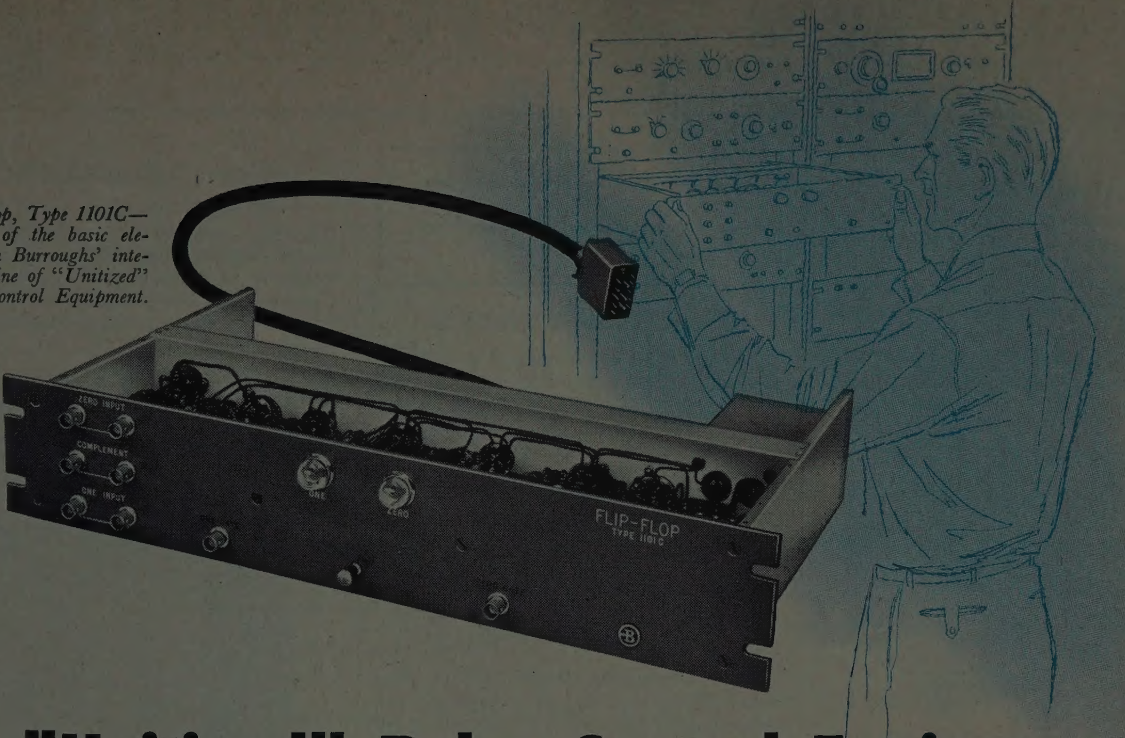
FOR APPLICATIONS IN RADAR, TELEMETERING, DIGITAL COMPUTING, TELEVISION, NUCLEAR PHYSICS



Flip-Flop, Type 1101C—another of the basic elements in Burroughs' integrated line of "Unitized" Pulse Control Equipment.

"Unitized" Pulse Control Equipment

saves time and money in electronic engineering

There's no longer any need to tie up engineering personnel with the time-consuming work of developing and "breadboarding" electronic test circuits. Burroughs, a leader in the office machine industry, now offers an integrated line of "Unitized" Pulse Control equipment covering all the basic functions in pulse circuit engineering. These one-basic-function units are designed with a maximum of flexibility to be used as building blocks for test systems ranging from the very simple to the most complex. Engineers need only make a block diagram of the apparatus needed, assemble the necessary Burroughs units in the plug-in rack, and interconnect them with the various standard coaxial cables and accessories. It's really that easy! It's equally easy to reassemble your units for a different project when your present tests are completed.

YOU SIMPLY "PLUG IN" BURROUGHS FLIP-FLOPS

Burroughs Flip-Flop, Type 1101C, demonstrates the one-basic-function principle that makes Burroughs "Unitized" Equipment so suitable for your needs.

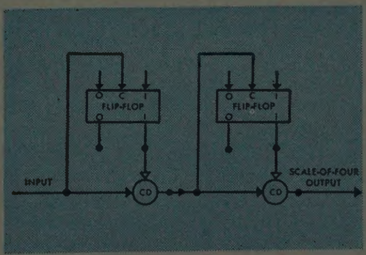
This flip-flop is a bistable circuit designed specifically to provide an output gating voltage to be used in coincidence circuits. The unit contains a pentode Eccles-Jordan circuit capable of being switched at rates up to 2.5 megacycles per second, with 0.1 microsecond pulses.

There are three inputs—Zero, One and Complement—operating from pulse amplitudes of 12 volts or more. Coaxial output jacks marked "Zero Gate" and "One Gate" supply either 0 volts or -23 volts at an impedence level of approximately 680 ohms.

Two neon lights on the front of the panel indicate the position of the flip-flop. A terminal block on the rear of the unit can be used to operate indicator lights installed at a remote point for visual monitoring.

Proved by more than two years of constant use, Burroughs "Unitized" Pulse Control equipment has been purchased by many leading electronic research organizations. Some of the users are: Massachusetts Institute of Technology, University of Michigan, Stanford Research Institute and National Union Radio Corporation.

Scale-of-Four Binary Counter Using Burroughs "Unitized" Equipment



The left flip-flop, Type 1101C, changes state with each input pulse, so that the left coincidence detector (CD) or gate, Type 1201B, is alternately opened and closed with succeeding input pulses, with the result that every other input pulse passes through the left coincidence detector, giving a count of 2. A similar flip-flop and gate combination cascaded to the first combination gives a total scale of 2 x 2 = 4. The number of flip-flop and coincidence detector combinations that can be cascaded is unlimited.

For full information on Burroughs "Unitized" Pulse Control Equipment, write or call Department 14B, Electronic Instruments Division, Burroughs Corporation, 511 N. Broad St., Philadelphia 23, Pa.

PULSE GENERATORS
COINCIDENCE DETECTORS
PULSE DELAYS
FLIP-FLOPS
PULSE GATERS
CHANNEL SELECTORS
MIXERS

ELECTRONIC INSTRUMENTS DIVISION
Burroughs
THE BEST KNOWN NAME IN OFFICE MACHINES

(Continued from page 21A)

Frequency Converter

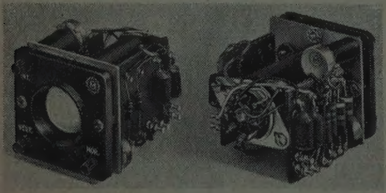
Avion Instrument Corp., Div. American Car and Foundry Co., 299 State Highway #17, Paramus, N. J., announces the development of a new frequency converter, Model 400. This portable 400 cps power supply operates from a standard 60 cps line and is capable of delivering up to 100 volt-amperes. It is suitable for laboratory use in the development of low-power aircraft equipment.

The output voltage is variable by a front panel control, from 105 to 130 volts, with regulation better than ±1 per cent. Frequency is adjustable from 380 to 420 cps with less than ±1 cps drift. Total harmonic distortion is less than 3 per cent. All these characteristics are independent of the power factor.

The basic components of this converter consist of a standard high-power audio amplifier, an adjustable tuned circuit, and an adjustable bridge circuit which controls the voltage output. The circuit contains a total of eight tubes and presents no unusual servicing problem.

Panel Mounting Oscilloscope

The use of miniature panel-mounting cathode-ray oscilloscope for instrumentation, in place of the conventional "pointer type" moving-coil meter, is made practical by the new 1 inch oscilloscope just developed for this purpose by the **James Millen Manufacturing Co.,** 150 Exchange St., Walden, Mass.

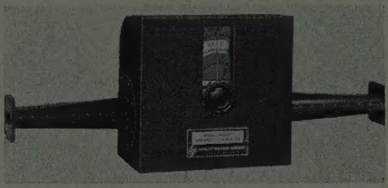


The new Millen instrumentation scope has a panel bezel matching in size and style the standard 2 inch square case meters. It uses the 1 Cpl one inch cathode-ray tube. This compact tube has a bright and sharp trace. There are many places in receiver, transmitter, amplifier and radar circuits where the wave shape and pulse form are of equal or even greater importance than magnitude. By the use of this new unit magnitude, phase displacement, wave shape, and so forth, are constantly visible on the scope screen.

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

Wave Guide Attenuator

A new kind of waveguide attenuator that gives direct and precise readings from zero to 50 db has been developed by the **Hewlett-Packard Co.,** 395 Page Mill Rd., Palto Alto, Calif.



The instrument, Model X382A, provides db readings that are completely independent of frequency. Phase shift is independent of attenuation setting and accuracy is within ±2 per cent of decibel reading. The attenuator is directly set and read, and requires no interpolation or charts.

Design of the instrument is unique in that the attenuation depends on the angular position of the attenuating film rather than its specific resistivity. Model X382A employs three resistive films, two mounted in line within each waveguide extension, and a third film rotatable axially in the center section. With all three films aligned in the same plane, there is no attenuation. Rotating the center film with the front panel knob increases attenuation proportionally to the cosine squared of the angle of rotation.

Model X382A is now offered only for X-band use at frequencies from 8,200 to to 12,400 mc. Calibrated range is 0 to 50 db, vswr is less than 1.15 throughout range and zero setting attenuation is less than 1 db. Power may be fed to either end. Model X382A is priced at $250.00 f.o.b. factory.

Varistors

A new line of Varistors (non-linear resistors) has been introduced by the **International Resistance Co.,** Special Products Div., 401 N. Broad St., Philadelphia 8, Pa.

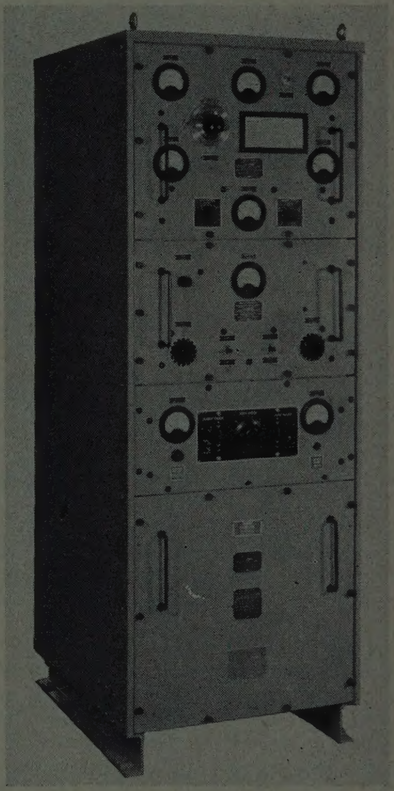
These units have many applications in circuits where sharp variation of resistance with applied voltage is required, and are available in 5 cell sizes in a wide variety of enclosures.

Designed to conform with MIL and JAN Specifications on humidity, shock, vibration, temperature cycling, solder pot, fungus resistance, and so forth IRC Varistors have low shunt capacitance and can be used effectively in rf circuits. They are claimed to have instantaneous response.

For further information write International Resistance Co. for catalog Data Bulletin SR-3.

Beacon Transmitter

Gates Radio Co., Quincy, Ill., has announced a new LF Homing-Beacon transmitter incorporating variable output, and available in a variety of combinations. Carrier power is adjustable from 25 to 400 watts by a front panel controlled autotransformer.



The Gates MO-3975 is available with or without remote control. High level voice modulation is employed, incorporating a peak-limiter to prevent over-modulation. Type 36 automatic code keyer operates from a phase-shift oscillator at 1,020 cps.

MO-3975 can be supplied for 51½ ohm output or a special antenna coupler is available to couple directly to an L or T type antenna.

Fully JAN approved, the transmitter is designed to withstand "Class B" shock tests and operate in ambient temperatures between −54°C and +65°C. Pressurized forced air cooling is employed.

Each section is mounted on a slide type drawer to provide easy access for maintenance.

Technical specifications of the MO-3975 include the following: Carrier range, 200–800 kc; Frequency stability, 0.005 per cent; Audio response, ±2 db 400–3,000 cps; Distortion, 6 per cent or less, 400–3,000 cps; Noise, 40 db below 100 per cent modulation at 400 watts; Power consumption, 2200 watts 115 v ac single phase at full power output.

Complete information may be had by writing the manufacturer.

(Continued on page 48A)

PROCEEDINGS OF THE I.R.E.®

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The Institute of Radio Engineers, Inc.

VOLUME 41 — *August, 1953* — NUMBER 8

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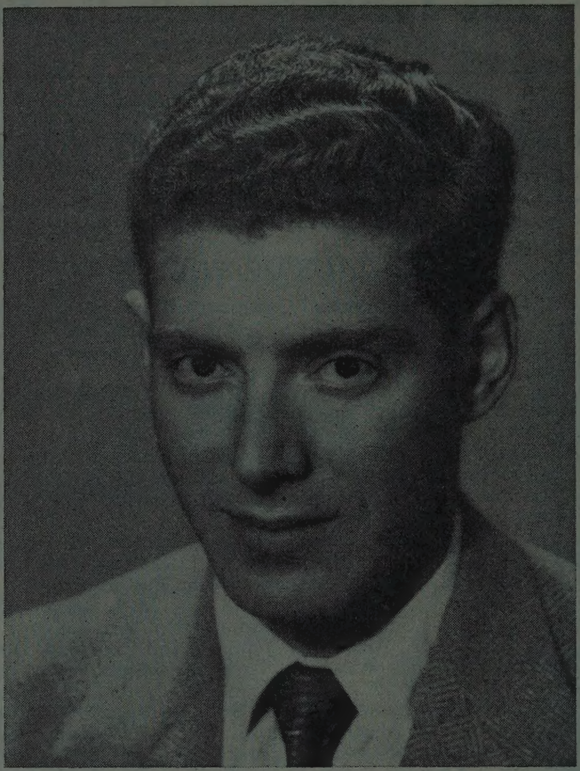

Amor L. Lane

CHAIRMAN, 1953

PROFESSIONAL GROUP ON ULTRASONICS ENGINEERING

Amor L. Lane, first chairman of the new Professional Group on Ultrasonics Engineering, was born on February 16, 1926, in the state of Pennsylvania. He received the B.S. degree in electrical engineering from the University of Pennsylvania in 1949.

From 1947 to 1949 Mr. Lane served in the United States Navy as an electronic technician's mate, interrupting his college studies.

Mr. Lane joined the electronic guidance section of the Guided Missiles Division of the Naval Ordnance Laboratory in 1949. In 1950 he transferred to the ultrasonic transducers section in the Underwater Ordnance Department. He became a group head in 1951 and section chief in 1953. This section is responsible for applied research, design, development, construction, and evaluation of electroacoustic transducers, with emphasis on polycrystalline ceramics.

For his work in underwater sound Mr. Lane received the Navy Department's Civilian Meritorious Award last January.

Mr. Lane joined the Institute as an Associate in 1949. During the past year he has been instrumental in founding the Professional Group on Ultrasonics Engineering. He was elected its first chairman in May of this year.

Mr. Lane is a member of the Acoustical Society of America, Tau Beta Pi, Eta Kappa Nu, and Sigma Tau.

Ultrasonics Engineering

A. L. LANE

As science and technology evolve, new and important fields of endeavor spring into being and assume their rightful place among their older brethren. What were little known of detailed principles become the bases of developed and expanded systems of thought which soon find their practical expression in humanly helpful engineering methods and equipment.

Ultrasonics in engineering is a striking example of this trend. Its development has been rapid and its applications many and increasingly valuable. It is therefore fitting that an IRE Professional Group on Ultrasonics Engineering should have been established, and that its interesting scope and encouraging plans should be described below by its Chairman, who is Section Chief, Ultrasonic Transducers Section, Underwater Ordnance Department of the Naval Ordnance Laboratories, White Oak, Maryland.—*The Editor.*

Ultrasonics has been called correctly the science of a coming technology. Only a few years ago the scope of ultrasonics was limited and the applications comparatively few. However, if the tremendous advances that have recently been made in ultrasonics are used as a measuring stick, the future promises even more astounding developments. There can be little doubt that a potent and versatile tool is emerging, which, even now, is proving extremely useful to the scientist and engineer both in the laboratory and in industry.

At present the investigation and application of ultrasonics embraces many diverse fields. In the laboratory ultrasonic techniques are used for studying specific heats, molecular processes, shear viscosity, and elasticity of liquids, the elastic and dissipative properties of solids, thermal relaxation phenomena, velocity, and attenuation in gases. Marine applications include ultrasonic depth indicators, underwater object and fish locators. In the medical field ultrasonic diagnostic and therapeutic devices are being investigated and developed. These include diathermy instruments, tumor locators, dental caries locators, and even a device to replace the dentist's drill. In industry ultrasonics is finding applications in wide variety. Nondestructive testing of materials, acceleration of chemical reactions, emulsification, coagulation, sterilization are a few uses in this area. Ultrasonic delay lines and electromechanical filters are being used in radio, radar, and digital computers. There may one day be ultrasonic washing machines.

In view of these tremendous strides, there has been a large and spontaneous demand from widely divergent fields, interested in or utilizing the generation and effects of ultrasonics, for an organization representing their interest. In response to this widespread demand the Ultrasonics Engineering Professional Group of the IRE has been organized. This Group is now taking its place among the other Professional Groups of the IRE, according to the IRE policy of decentralization into smaller, more compact Groups formed on the basis of professional interest.

The newly formed Ultrasonics Engineering Group will benefit both the ultrasonic engineer and industry as a whole. This will result from an attempt to bring together the literature on ultrasonic applications and associated circuitry which heretofore has been dispersed in a wide variety of publications. Both the TRANSACTIONS of the Ultrasonics Engineering Group and the PROCEEDINGS OF THE I.R.E may be used for this purpose. Moreover, the TRANSACTIONS often will enable more rapid publication of certain papers. In addition to the TRANSACTIONS, the Ultrasonics Engineering Group will sponsor symposia and local meetings which will provide ultrasonic engineers with their own outlet of expression. Industry will be helped, since the interchange of ultrasonics information will be available to the entire IRE membership of 30,000 engineers and scientists, as well as the additional thousands of readers of IRE literature.

An Evaluation of the I.R.E. Professional Group Plan*

W. R. HEWLETT†, FELLOW, IRE

Summary—**The rapid growth during the past ten years of the radio engineering profession resulted in the initiation by the I.R.E. of the Professional Group Plan. The significant steps in the development of the Professional Groups are reviewed and the value of the Professional Group Plan to the Institute and to the Profession as a whole is evaluated. The importance of realistic consideration of problems arising from the group expansion program is stressed.**

THE YEAR 1953 saw 40 of the 43 technical sessions at the National Convention sponsored by Professional Groups. This is indicative of a change that has been gradually taking place in the Institute since the Professional Group plan was first initiated. It is the purpose of this article to review briefly some of the factors that led up to the Professional Group system and to indicate the advantages that will accrue to the membership as a result.

A little less than five years ago the Professional Group plan was initiated within the I.R.E. This plan consists of a voluntary association of Institute members who are interested in seeing that their particular branch of the electronic industry is adequately covered by Institute activities. Upon approval of the Institute and after meeting certain formal requirements, a Professional Group may be set up to insure this aim. In this short five-year period, the plan has grown until there are now no less than 19 Professional Groups with a paid-up membership of approximately 12,500.

To go back a bit in history, the Institute was founded in 1913 by a small group of engineers and scientists who were interested in or directly concerned with the infant field of "radio engineering." It is interesting to look at the type of problems that concerned this early group. Selecting a few papers from the first two years of PROCEEDINGS publication, one finds articles by all three of our founders, two of whom we have honored today. These papers treated such subjects as: "Radio Operations of Steamship Companies," "The Seibt Direct Indicating Wave Meter," "The Heterodyne Receiving System," papers on "Experimental Tests of the Radiation Law of Antennas" by Michael Pupin, "The Daylight Effect in Radio Telegraphy" by Kennelly, "The Audion-Detector and Amplifier" by Lee deForest, and "Radio Traffic" by David Sarnoff.

Although the subject matter of any of these papers might make it the legitimate claim of a modern Professional Group, the basic difference is that these papers were of paramount interest to all "Radio Engineers" at that time, and not just a special group.

In contrast, these typical papers appeared in last year's PROCEEDINGS: "Network Synthesis by the Use of Potential Analogs," "The Binac," "A Homo-Polar Tachometer for Servo-mechanism Applications," "Transistor Forming Effects in N Type Germanium."

These papers were selected to represent the high degree of sophistication the art has assumed.

Several basic factors have contributed to the broadened scope of interest of the I.R.E. membership. About 1940, the profession of radio engineering was pretty well confined to the field of communications and the directly supporting activities. The war had a tremendous impact on the entire field, and electronics emerged from the war as a basic tool of industry as a whole and not just the bread and butter item of a small group of engineers in the field of communications. The extent of this expansion is perhaps best indicated by the fact that for the 12 years prior to 1940 the Institute membership had been almost constant at 6,000, but starting in 1940, the annual increase in membership has averaged close to 2,500 per year.

Dr. William Everitt[1] stressed electronics as interested primarily in extending man's senses, and proposed this definition:

> "Electronics is the science and technology which deals primarily with the supplementing of man's senses and his brain power by devices which collect and process information, transmit it to the point needed, and there either control machines or present the processed information to human beings for their direct use."

An acceptance of this definition and a tacit assumption that the I.R.E. is the professional society primarily responsible for the field of electronics immediately indicates the enormous scope of the I.R.E.'s responsibility. The diversity of interests represented by this definition obviously offers a tremendous challenge to the Institute, a challenge that must be accepted if the Institute is to live up to its basic objectives as set forth so well in its constitution, here quoted.

> "Its objectives shall be scientific, literary, and educational. Its aims shall include advancement of the theory and practice of radio and allied branches of engineering and of the related arts and sciences, their application to human needs, and the maintenance of a high professional standing among its members."

The Institute's answer to this challenge has been the Professional Group system. It is interesting to note that the author of the broad definition of electronics, Dr. Everitt, together with Dr. Raymond Heising, played a leading role in the establishment of the Professional Group system. As previously noted, the Professional Group plan was set up so that the impetus for the formation of a Group must come directly from the membership, not from some arbitrary wish on the part of Institute headquarters. Operating on this basis, it became evident by 1951 that, not only was the Professional Group plan gaining rapid support from the membership, but also, that if the best interest of the profession were considered, even greater stimulation must be given to the movement. This step was taken by a strong statement of policy by the Board favoring decentralization and the adoption of a plan whereby publication of technical and specialized papers by the Professional Groups would be stimulated.

This plan provided that, on a one-time basis, the Institute would match, up to $1,000, any sums raised by a group either through assessment or by the holding of symposia. This plan has since been expanded so that the Professional Groups may count on continued financial support from the Institute, with the funds to be furnished on a matching basis subject to certain restrictions as to maximum amount. The result of the encouragement given the Professional Groups by the Institute, both financial and moral, resulted in a noticeable strengthening of the Group plan. In 1952, for example, 22 separate TRANSACTIONS were published by Professional Groups. It is interesting to note that, although responsibility for obtaining subject material for these TRANSACTIONS rests primarily with these groups, a very considerable support is furnished by the Editorial Staff of the Institute. The increased vigor of the Professional Groups is not only indicated in their publications but also in the symposia they sponsor. Last year there were about 12 specialized symposia held throughout the country that were either sponsored or cosponsored by Professional Groups.

The flow of assistance, however, has not been a one-sided proposition. More and more the Institute is looking to the Profession Groups for aid. In the case of publications this help has been two-fold—one direct and one indirect. The Editorial Department looks to a Professional Group for definitive papers in the particular field of interest of that Group and has so streamlined its editorial procedure that papers recommended by the Professional Groups now receive preferential treatment. The indirect benefit has been that the PROCEEDINGS is relieved of the very considerable pressure to publish detailed and often specialized papers in various fields now covered by Professional Groups. In many instances these papers can be more satisfactorily handled through Group TRANSACTIONS. This permits the PROCEEDINGS to concentrate on papers of a more basic nature and of greater general interest to the Institute as a whole.

In the national conventions the Groups are playing an ever increasing rôle, as witnessed by this year's national convention. The same thing to a lesser extent is true of conventions held elsewhere in the United States. For example, at the West Coast Convention held in Long Beach last year 14 groups sponsored technical sessions. Even the Sections themselves are finding the local

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† Hewlett-Packard Co., Palo Alto, California.

[1] W. L. Everitt, "Let us re-define electronics," PROC. I.R.E., vol. 40, p. 899; August, 1952.

Professional Group Chapters of assistance and often will request that a Group Chapter sponsor one or more meetings during the year for the general membership.

We are now in a position to stand back and evaluate our Professional Group system. Most significant, of course, has been the recognition by the Institute that if the best interests of its members and the profession as a whole are to be preserved it must adopt some sort of division along professional lines. The requirement for a division of this type indicates the profession has become sufficiently complex that it can no longer be served properly by media common to all members. Although there are large fields of common interests which still should and will be handled by the present media, the specialist must be provided with means for the direct and rapid interchange of information within his own field. These means could not be provided on an economical basis by the previously existing publications, conventions and meetings.

A less obvious benefit of the Group system has been to broaden the base of membership participation. This in itself is a fact of no mean significance. Examples of this are to be seen in the greater opportunity among our membership for experience in organization and leadership. Many of the men so trained will eventually work into Institute management where their reservoir of experience will be of great value. Additional opportunities have been offered in the field of publication and rapid dissemination of technical papers and again this will be of value to the Institute. Through the ever increasing number of conferences and symposia in which the Professional Groups are playing a leading role, the membership of the Institute as a whole will derive great benefit. In addition, many interesting experiments on how to organize and run a conference are being tried by the groups. Some of the more successful will undoubtedly have a direct effect on our national convention.

In many ways the Institute might be likened to a parent with 19 vigorous and growing children. Most of these children have been carried through their first few critical years and have survived the usual childhood diseases of a very young organization. Most of the Groups are now sturdy children rapidly approaching maturity. Already the Institute is leaning heavily on them. There are, however, critical periods ahead, both for the Institute and for the Professional Groups themselves. New organizations are usually started by a few energetic people who through their enthusiasm and drive establish the organization as a going concern. However, in many cases the interest of these people is transitory, either because they look for new worlds to conquer or through general lack of interest once the basic problems of organization have been surmounted. It is, therefore, imperative that the Professional Groups insure that young and vigorous personnel are attracted into the management and guidance of the Group. When the time comes for the original organizers to step aside there will be a well-trained and capable team to insure the continued good management of the Group. On the other hand, the Institute itself must furnish intelligent leadership to its Professional Group children. It will not be long before these Groups are fully grown adults and as such will ask and must be given an important role in the running of the Institute. Institute leadership must be such that the first allegiance of the Professional Group will be to the Institute as a whole.

We are probably entering one of the most critical phases in the development of our Group system. We have gone too far to turn back, and must make the system work or see the Institute shatter into small pieces, the total of which would never be as strong as the whole, nor as capable of serving the interests of the profession.

The Groups have become large enough and sufficiently well defined that basic questions are arising with respect to the relation of Groups to each other and of Groups to the Institute itself. It is imperative that these and similar problems as they arise be faced squarely and considered honestly by the Professional Groups and the Institute alike. On the soundness of these decisions rests the long range success or failure of the Professional Group plan. The basic yardstick must be "What is best for the Institute and the profession as a whole?" If this yardstick is conscientiously used, our Professional Groups will thrive, membership in the Institute will continue to rise, the service performed to our members and profession will grow better, and the Institute of Radio Engineers and its Professional Group system will be a model of how a professional society can meet the demanding requirements of an ever-increasing technical field.

Is Complexity of Military Electronics Necessary?*

R. J. NORDLUND†, SENIOR MEMBER, IRE

Until recent years, adequate attention had not been paid to the engineering of systems embodying both complex electronic mechanisms *and* the human operator of such mechanisms. When such composite systems are thoughtfully designed, they often turn out to be highly complex. Complexity may bring unreliability in its train. However, painstaking planning and pretesting of such systems may restore reliability. The following paper comprehensively analyzes these situations and is accordingly a useful guide to engineers attacking such problems.—*The Editor.*

Summary—**An analysis is made of the reasons for the trend of military electronics toward more system complexity. Some interesting aspects of the relation of complexity to unreliability are investigated, and a distinction is established between circuit complexity and operation complexity. The soundness of the choice of complexity is established by curves of a general "trend" nature. Systems engineering is proposed as a solution to the complexity versus reliability problem, and as a challenge to the young engineer.**

MANY PEOPLE believe that the military departments have "gone off the deep end" in the matter of system complexity. Statements have been made, in the press and elsewhere, to the effect that we are soon going to attain "push-button" warfare. This seems to imply the replacement of the human as part of the basic system. Certain of the more intricate circuits of the computation type are rather thoughtlessly called a "brain."

I think you will agree that the most important contribution of the human in a servo loop, in a dynamic situation, is his ability to reason and correct a wrong trend or event if and when it occurs. With most humans, this is true only so long as events occur at a reasonable rate, so that confusion does not ensue, and emotional stability is not upset. On the other hand, electronic instrumentation can be devised which can perform certain functions at an extremely rapid rate, without confusion. Even a perfunctory analysis of almost any branch of air warfare immediately shows that one of the inherent ingredients of any such battle is an ever-changing situation to which reasoning power can be profitably applied. It immediately becomes apparent that an optimum combination of human and instrumentation must exist for the most precise solution of any given dynamic problem. The faster and higher an airplane flies, the more necessary this becomes; in fact automaticity, and hence, complexity, will *have to* increase in proportion to the rate of increase of speed and altitude.

World War II saw the first application of complex electronic gear to the air battle (as well as to sea and land warfare). The introduction of radar to extend human capabilities in detection and ranging for fire control and bombing was certainly a milestone both in the military field and in the electronic industry. While it is undoubtedly true that this development contributed greatly to the achievement of mastery over the enemy, it is also true that specific results fell far, far short of the goals envisaged by

* Decimal classification: R560. Original manuscript received by the Institute, April 2, 1953.

† Bombing Consultant, Armament Laboratory, Wright Air Development Center, Dayton, Ohio.

equipment designer and manufacturer. Two basic reasons for this are unreliability and complexity of operation.

In retrospect, it is apparent that this unreliability was inevitable, considering the astonishingly short period given to development and design. But the greatest reason for its unreliability was the revolutionary nature of the impact of such complex gear, introduced in such a short time, upon the maintenance organizations of the services. This was like asking a building construction electrician to repair a television set and withholding the circuit diagram. Looking back, it seems a wonder that any effectiveness was achieved at all, and much credit must be given to the electronics industry and the cooperative spirit and understanding of its technical representatives in the field.

The other reason was termed "complexity of operation." It will be shown that this is an entirely different thing than circuit complexity, and this is very important. Complexity of operation brings in the human who uses the equipment. The electronic equipment used in the Air Force in World War II simply required of the user so many parallel actions and this at a time of great emotional strain, being under attack, that his ability to reason and rationalize was destroyed. The result was a great plague of gross errors—in the case of bombing—which was difficult for the designer of the equipment to understand or explain. These gross errors were the result of the operator actually forgetting to perform some important calculation, or actually making an erroneous calculation, or failing to apply reason to a sudden turn of events.

In the developmental evolution of Air Force electronic equipment, during and immediately following World War II, analysis clearly showed what the limitations of the user were. An important decision had to be made. This decision was to accept the limitations of the human and to stay within these limitations by complex instrumentation which accomplished many of the tasks and computations automatically. The aim was to achieve maximum utilization of the reasoning capability of the human, which cannot be instrumented, by removing the confusion factor caused by overloading him. The penalties to be incurred by circuit complexity in the form of unreliability and difficult maintenance were clearly understood and considered in the decision.

There seems to exist a certain amount of confusion between the terms circuit complexity, operating complexity and reliability. It is obvious that circuit complexity and reliability are basic problems facing the equipment designer. The problem is to accept the fact that circuit complexity is necessary and to go about the business of achieving reliability in spite of complexity. A certain amount of introspection can convince one that, in the eyes of the user, circuit complexity and unreliability are synonymous. If a black box performs its function day after day, never requiring disassembly or attention, to the user it is a simple gadget regardless of the wondrous jumble of tubes, capacitors, resistors and whatnot which may be hidden inside.

The goal of the designer, then, is to make the user believe that a complex equipment is really simple. We have many examples of this in our homes. Almost every home in the country has a telephone. To millions of persons this is assuredly a simple device, since it never fails and is easy to operate. To those who spent thousands of hours designing, testing, redesigning and retesting to achieve this almost perfect reliability, it was hardly a simple task, nor is the welter of electronic instruments in the modern automatic exchange to be considered simple. The telephone system, incidentally, also illustrates to some extent the earlier contention that *from the system standpoint*, in both the economical and accuracy aspects, optimum utilization should be made of the human being *and* electronic instrumentation. The telephone system still utilizes people to apply reasoning power to sort out the weird questions which we can pose from the other end of the line, but has removed from the human the mechanical tasks of making proper connections to connect us to the right party. Many other examples in daily life need not be mentioned here.

An attempt will now be made to demonstrate by means of some simple curves that the decision to try to achieve proper balance between human and instrumentation, at the cost of circuit complexity, was a most wise choice on the part of the Air Force.

Fig. 1 contains five curves based on factual data concerning one particular type of Air Force weapon system. Because of this factual nature, the figure is lacking in a quantitative sense, since these data are classified information. Nevertheless, much can be learned from a study of these curves. The curve entitled "Instrument Accuracy" is included because precision of computational accuracy does have a bearing on complexity and reliability, and the disparity between instrumental and operational accuracy is explained by the other three curves. It should be noted that a high order of instrumental accuracy was available even at the beginning of World War II. However, the disparity between this and actual operational accuracy was enormous due to large demands upon the human at the wrong time.

Introduction of radar increased the operator's capability but complicated his task to a great degree so we have the "operator's task" curve rising to a broad peak around 1944, with a corresponding degradation in the "operational reliability" curve. "Instrument Accuracy" was also degraded because of the coarser nature of the radar-data inputs and of the lesser accuracy of the electronic solution versus the previous mechanical computation. These curves improved slightly between 1944 and 1946 reflecting increasing capability in maintenance and adjustment and admirable support of the Air Force by the electronics industry.

The first radical change in these curves coincides with the introduction of the new equipment designed to most effectively use the human operator. The "operator's task" curve goes rapidly from complex to relatively easy, and at the same time the "operational reliability" curve rapidly falls to a dangerously low level. The double low spot in the "operational reliability" curve resulted from the introduction, about a year apart, of two different configurations of this new concept, the second more elaborate in capability and complexity than the first. The climb back to respectability of the "reliability" curve was achieved by an agonizing and very expensive "bootstrap" operation in which basic mistakes of installation which adversely affected the equipment environment were corrected, the usual two year gap between engineering knowledge of deficiency and actual production-line quality was overcome, and the capability of the using organization made a surprising improvement. This was a period during which the Air Force was learning to live with circuit complexity. That this was possible is demonstrated by the rapid rise of the "reliability" curve during 1949 to 1953.

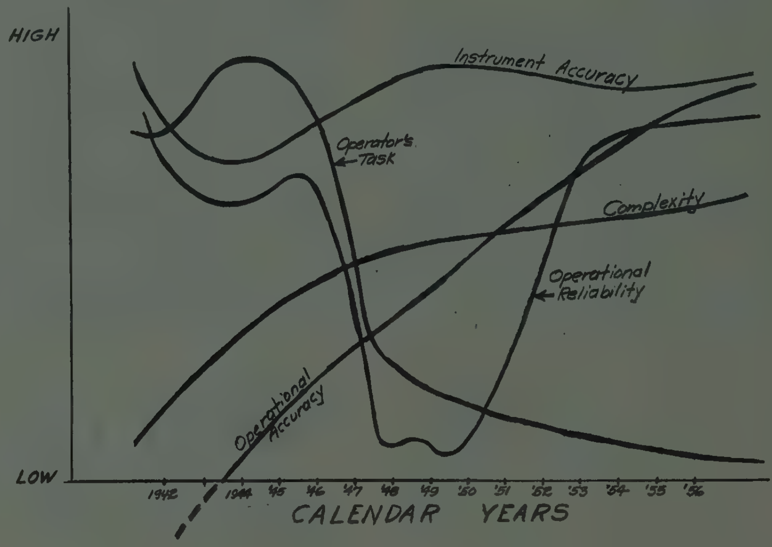

Fig. 1—USAF system concepts.

The "complexity" curve is a rather arbitrary one, since this is hard to measure in chart numbers. It does illustrate the revolutionary nature of what took place between 1942 and 1948 in this matter so important to the members of the Institute.

The significant thing about this figure is, of course, the "operational accuracy" curve. The fact that it shows a steady rise during all the gyrations of the other curves is of tremendous importance to the nation and, it is believed, proves that the decision in favor of circuit complexity was sound.

The fact remains, unfortunately, that reliability is at present far from acceptable, and while learning to live with electronic-equipment complexity we have continuously had present that ugly unwanted bedfellow "unreliability." We know now that we have capability, but for it we are paying far too high a price in maintenance, attrition and mission failures. The challenge obviously is to learn how to build reliability into complex electronic equipment. The Air Force has not the slightest doubt that the electronics industry will be able to achieve this. We are, of course, interested in how soon, not only because of cost considerations, but also because the quality of the nation's security is inextricably involved.

The Air Force technical personnel have recently come to recognize that the only way to achieve a position where our customers, the operating Air Forces, can be led to believe that complex electronic equipment is in fact simple and reliable is by approaching the problem from the complete "systems" viewpoint. This is how a "simple" black-rubber telephone instrument was achieved.

A modern weapon system is an extremely complicated thing and the problems of adequate technical administration of such a project are enough to stagger the imagination. The galaxy of specialized talent which must work together on such a project includes representation from all fields of the aerodynamic, mechanical, optical, geodetic and electronic arts, as well as human resources research, basic materials research, aeromedical research, not to mention the basic military sciences. The Air Forces' technical administration is keenly and humbly aware of the enormity of the task and looks to the industry for a large measure of the necessary aid.

The author believes that it is in systems engineering in complex electronics-systems work that the real challenge to the present day young engineer lies. It is work that does not appeal to the type of person who desires to specialize and acquire a great store of knowledge about some distinct and unique portion of the field. Nor does it continue for long to appeal to the type of individual who likes his days to be serene and his nights untroubled. It is difficult work, but unbelievably rewarding in the sense of real accomplishment and service. The young man who is fortunate enough to get in on the conception of a systems project, and can stay with it through the many years required for development of such a project, emerges with a mature sense of his own capabilities and is usually then equipped to engage in technical administration work. Systems engineering work is the most broadening of any phase of the profession and is a liberal education in itself.

In general, the Air Force has enjoyed the best of relations with the electronics industry, and is deeply appreciative of the spirit of understanding and cooperation that has existed. Air warfare is young and dynamic, and consequently subject to continuing change based on experience and technological advancement. Its technical programs are full of unique problems—unique because the environment in which airborne electronic equipment "lives" is certainly unique, and this is one of the most difficult problems to overcome in the struggle to achieve reliability in spite of complexity. These special problems are of the type which require the most thorough cooperation between the industry and the military departments.

Concluding points:

- (a) Modern military equipment, to achieve the most precise solution to dynamic problems, must rely on complex instrumentation to permit optimum use of the human being.
- (b) To the user, complexity of itself is of no real insignificance except as it affects reliability.
- (c) The "systems" concept in engineering administration and technical control is essential to insure success in achieving reliability.
- (d) Systems engineering is the new frontier in the engineering profession.
- (e) Our goal must be to achieve reliability in spite of complexity.

Large Reductions of VHF Transmission Loss and Fading by the Presence of a Mountain Obstacle in Beyond-Line-of-Sight Paths[*]

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Summary—**A detailed investigation of the probable mode of propagation in vhf operation over mountain obstacles has been made. Theory indicates that tremendous gains in received signal strength—above those obtained over a smooth spherical earth—may be expected from diffraction over an appropriate knife edge located in the path. Experimental verification of the principles involved is reported. In addition, the fading was found to be small and even essentially nonexistent on certain very long obstructed paths.**

VERY SUCCESSFUL VHF communications have been reported with essentially no fading over very long paths across mountainous terrain. The authors have just completed a detailed examination of these reports and visited engineers and scientists of the Mutual Telephone Company of Hawaii; the Electrical Communication Laboratory, Ministry of Communications, Tokyo, Japan (similar to the Bell Telephone Laboratories in the U. S. A.); the Radio Research Laboratories, Ministry of Postal Services, Tokyo, Japan (similar to Central Radio Propagation Laboratory in the U. S. A.); and the Civil Aeronautics Administration in Alaska.

In a preliminary study prepared in the Office of the Chief Signal Officer, Lt. Col. H. V. Cottony, USAR, on military leave of absence from the National Bureau of Standards, pointed out that a considerable increase in received signal strength would result when a large knife-edge obstacle is located at the midpoint of an 85-mc, 100-mile path. Japanese scientists have made similar theoretical investigations and have substantiated their findings by experiments.[1]

Fig. 1 shows the results of using the four-ray method of Schelleng, Burrows and Ferrell[2] for computing diffraction across a knife edge at radio frequencies. This figure shows the expected basic transmission loss[3] at 100 mc over transmission-path distances of 50 and 150 miles. The basic transmission loss is defined to be the ratio, expressed in decibels, between the power delivered to an isotropic transmitting antenna divided by the received signal power available from an isotropic receiving antenna. The actual transmission loss on a circuit may thus be obtained simply by subtracting the effective transmitting and receiving antenna gains, expressed in decibels related to an isotropic antenna, from the basic transmission loss shown in Fig. 1.

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[1] S. Matsuo, "The Method of Calculating VHF Field Intensity and Research on Its Variation," Report of the Electrical Communication Laboratory 621.39.001.6(047.3), Ministry of Telecommunications, Tokyo, Japan; August, 1950.

[2] J. C. Schelleng, C. R. Burrows and E. B. Ferrell, "Ultra short-wave propagation," PROC. I.R.E., vol. 21, pp. 427–463; March, 1933.

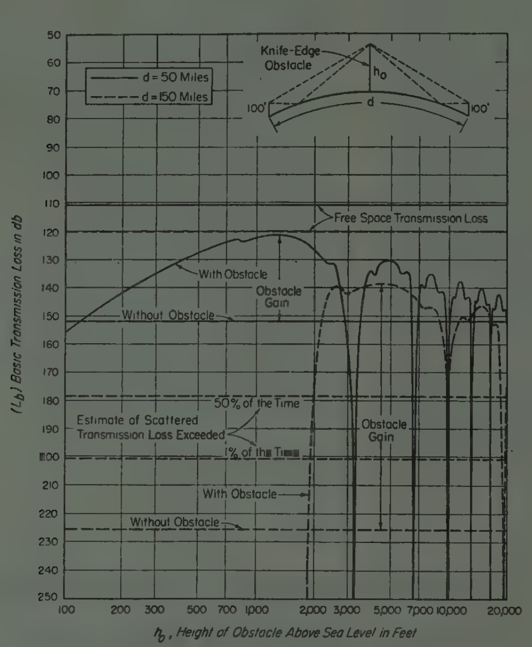

Fig. 1—Theoretical obstacle gains at 100 mc, assuming four-ray knife-edge diffraction theory. Transmitting and receiving antenna heights each 100 feet above the surface.

To simplify calculations, the ground-reflection coefficient has been assumed to be (−1), thus neglecting the small phase and amplitude changes expected for an actual earth, as well as the divergence of the energy reflected from the curved earth's surface. With this approximation, the height of the obstruction must be greater than the elevation of the common horizon to have an "obstacle gain"—as the computed transmission loss is infinite at this point, e.g., 1,850 feet for a 150-mile path. It may be seen from Fig. 1 that for particular combinations of antenna heights, obstacle heights, and frequency, tremendous "obstacle gains" in received field strength are to be expected. The "obstacle gain" is defined, as shown in Fig. 1, to be the ratio of the received signal power expected with and without the obstacle. Experimental evidence has been obtained of the existence of these very large predicted "obstacle gains" in field strength. These hitherto unexpected gains are being used in obtaining very satisfactory vhf radio communication over severely obstructed transmission paths in mountainous regions.

An example of the use of "obstacle gain" examined in detail can be found in the 38-mc 160-mile transmission circuit tested by the Civil Aeronautics Administration between Yakutat and Gustavous, Alaska. This path passes over the 8,000-foot level of Mt. Fairweather from two low terminals (approximately 200 feet) at the transmitting and receiving locations. The profile of this path, together with a typical sample signal-strength recording made by the Civil Aeronautics Administration on this circuit, is shown in Fig. 2.

On the basis of smooth-earth diffraction theory with the mountain removed, the calculated transmission loss for the 38-mc circuit would have been 207 db; however, on the assumption that Mt. Fairweather acts as a single 8,775-foot knife-edge ridge, calculations using the four-ray diffraction theory indicate that only 127-db transmission loss would be expected. The observed transmission loss was approximately 134 db, or within 7 db of the value indicated by the knife-edge diffraction calculations. These values correspond to a calculated "obstacle gain" of 80 db and a measured "obstacle gain" of 73 db greater than the field strength to be expected over a smooth spherical earth. Thus it appears that the mountain behaves effectively like a knife edge. Effective transmitting and receiving antenna heights of 50 feet and gains of 8 db over an isotropic antenna have been assumed in the above calculations, together with an allowance of 4 db for transmission-line losses.

Over a smooth spherical earth, signals arrive well below the radio horizon not only by the normal diffraction process around the curved surface of the earth but by scattering from the turbulent atmosphere. This turbulence is greatest near the surface of the earth. It is the turbulence of that portion of the higher atmosphere which is within line of sight of both transmitting and receiving antennas that is considered responsible for the relatively weak scattered, fading signal normally observed at distances beyond the horizon.[4] An estimate of the scattered-signal basic-transmission loss exceeded, for 1 per cent and 50 per cent of the time at 150 miles on 100 mc over a smooth spherical earth, is also shown in Fig. 1. This estimate was obtained by extrapolating the NBS Cheyenne Mountain data[4] so as to correspond to the 100-foot transmitting and receiving antenna heights of this example. It may be noted that the calculated "obstacle gain" of the properly situated knife-edge ridge reduces the transmission loss to such an extent that the received signal strength will far exceed the signal received by way of atmospheric scattering process, effectively eliminating fading arising from this scattering.

In addition, when an obstruction is situated at the proper location, and extending well above the horizon of both the transmitting and receiving antennas, the radio transmissions travel through the most turbulent and disturbed region of the atmosphere, close to the surface of the earth at a relatively large grazing angle. This large grazing angle reduces the possibility of the existence of numerous additional variable-transmission paths from the highly turbulent lower portions of the

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[3] K. A. Norton, "Transmission loss in radio propagation," Proc. I.R.E., vol. 41, pp. 146–152; January, 1953.

[4] J. W. Herbstreit, K. A. Norton, P. L. Rice and G. R. Shafer, "Radiowave Scattering in Tropospheric Propagation," I.R.E. 1953 Convention Record.

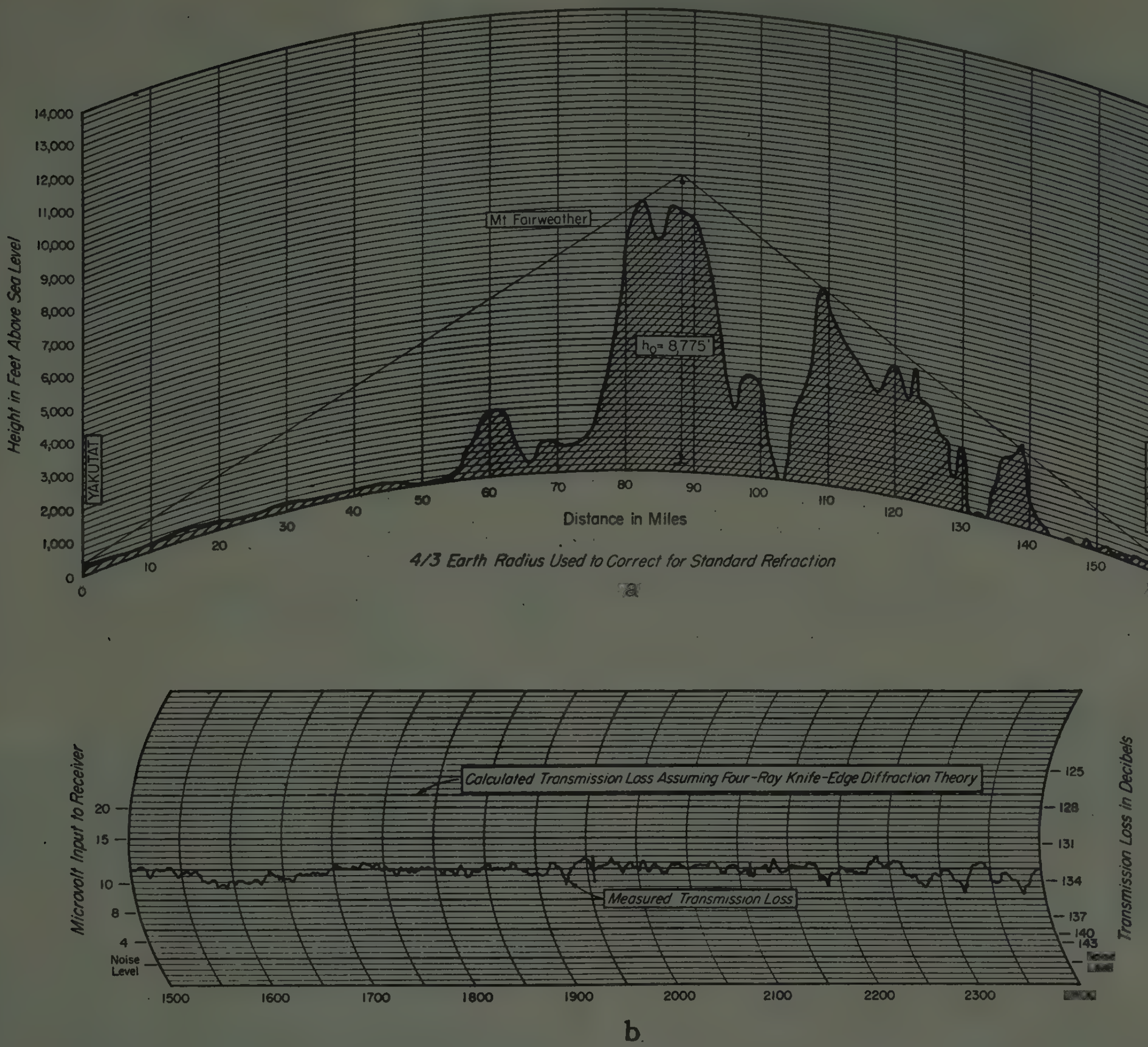

Fig. 2—(a) Profile of path from Yakutat to Gustavous, Alaska; $h_t=h_r=50$ feet; $f=38$ mc. (b) Typical sample recording on the above path; 50-watt transmitter power; $h_t=h_r=50$ feet. Local Alaska time, June 16, 1948.

atmosphere. In this way, the principal cause of tropospheric fading is effectively eliminated. The absence of severe tropospheric fading on a long-obstructed vhf transmission circuit is strikingly illustrated by the typical recording in Fig. 2. Over a period of 30 days' recordings on this path, the instantaneous transmission loss was found to vary from its mean value by less than ±2 db.

It is concluded that the hitherto considered disadvantageous high-mountain ridges in vhf transmission paths can, in fact, be of tremendous advantage in reducing both transmission loss and tropospheric fading. Future vhf siting practices in mountainous regions should be modified to take advantage of "obstacle gains" in accordance with the principles outlined above.

Low Level Synchronous Mixing*

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ABSTRACT

A SYNCHRONOUS DETECTION SYSTEM using same oscillator to furnish power for two signal channels has been used to obtain sensitivities on order of 110 dbm at 100 kc, 9 kmc, and 33 kmc.

The system operates by channeling the energy from an unmodulated RF source into two isolated paths. A portion of the energy is phase or amplitude modulated in one path termed the "signal channel." The remainder of unmodulated carrier is directed through the other channel, the "reference channel." The amplitude and phase of the energy in the reference channel is carefully controlled. The power output of each channel is mixed in a detecting element, and the modulation component is amplified and metered.

The behavior of synchronous mixing with amplitude modulation is predicted from a mathematical analysis based on a series expansion of the volt-ampere characteristic of the nonlinear detecting element. This analysis shows that the modulation component appearing at the output depends

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† The Johns Hopkins University Radiation Laboratory, Baltimore, Md.

upon the product of the reference power and the signal power. In addition, the modulation component is a linear function of the signal power. Further analysis with the series expansion was carried out for the case where the signal power is harmonically related to the reference power. In this situation, the theoretical sensitivity is less than in the fundamental case but is still appreciable.

When the signal channel is phase modulated, a simpler analysis is employed based upon the Fourier series expansion of the transconductance of the detector. This analysis leads to the same results as the amplitude-modulated case except that the second harmonic of the modulating frequency may be optimized in the output by properly adjusting the phase difference between the two channels.

The results of the theoretical analysis have been verified experimentally at three different frequencies. A low frequency, 100 kc, was chosen to obtain information on harmonically related energy and to verify the results of the phase modulated case. At this frequency it is possible to obtain phase modulation with very little accompanying amplitude modulation. A high frequency, 33 kmc, was employed primarily to examine the possible use of low-level synchronous detection for use in detecting 99-kmc power.

The experimental results showed that the sensitivity of the synchronous detector as compared to the crystal-video detector was approximately 30 db greater. Furthermore, the experiments demonstrated that the "law" of the detector was a linear function of signal power. For maximum sensitivity a pronounced optimum reference power, −20 dbm at 9 kmc and −40 dbm at 24 kmc was observed. Harmonic detection of 3-mm waves with 6-mm energy in the reference channel exhibited a 10-db increase in sensitivity as compared to crystal-video detection. Experiments were also made to examine the relationship between sensitivity and the modulating frequency.

The principal advantages of synchronous mixing are the sensitivity and dynamic range which can be attained with a minimum of electronic equipment. A sensitivity comparable to that of a superheterodyne receiver and a dynamic range greater than 60 db was easily attained. The disadvantage of synchronous mixing lies in the need for a reference-power connection between the source and the detector.

Unipolar "Field-Effect" Transistor*

G. C. DACEY† AND I. M. ROSS†

Summary—**Unipolar "field-effect" transistors of a type suggested by W. Shockley have been constructed and tested. The idealized theory of Shockley has been extended to cover the actual geometries involved, and design nomographs are presented. It is found that these structures can be designed in such a way as to yield a negative resistance at the input terminals. The characteristics of several units are presented and analyzed. It is shown that these characteristics are in substantial agreement with the extended theory. Finally a speculative evaluation of the possible future applications of field effect transistors is made.**

PART I, THEORY AND DESIGN

1. *Introduction*

THE "FIELD-EFFECT" transistor was first proposed by W. Shockley, and experiments of G. L. Pearson and W. T. Read gave some hope for ultimate success. Since that time, certain structural modifications have been proposed and the theory has been considerably enlarged. A detailed theoretical exposition has been prepared by Shockley.[1] We shall borrow liberally from this paper.

It has now become possible, through the advance of technique, to make units which are structurally similar to the ideal model of the theory, and to obtain a quantitative verification of the theory. In this paper we shall first review the ideal theory from a physical viewpoint giving plausibility arguments rather than proofs. We shall then consider certain modifications of the ideal theory suggested by the experiments and shall obtain design charts. The experimental verification of the theory will be given in some detail and finally relative merits of "field-effect" structures will be discussed.

2. *Shockley's Ideal Theory*

In essence, a field-effect transistor can be regarded as a structure containing a semiconducting current path, the conductivity of which is modulated by the application of a transverse electric field. In particular, we shall consider the structure shown in Fig. 1. To fix our ideas, let us suppose that the central portion of the unit is made of *n*-type germanium and that two *p*-type regions are centrally located as shown in the figure. We shall call these regions the *gate* for reasons which will presently become clear. In addition we suppose that ohmic contacts have been made at both ends. The contact at which the carriers flow into the germanium is called the *source* and the contact at which they flow out is called the *drain*.

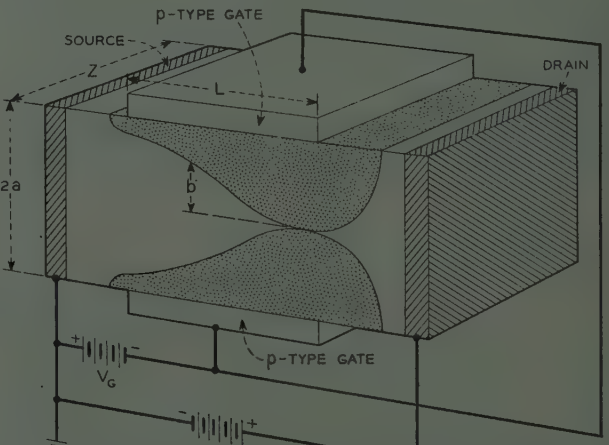

Fig. 1—Schematic diagram of field-effect transistor.

Now suppose that the gate is connected to the grounded source and that the drain is made slightly positive. Under these conditions current will flow from drain to source. This current consists of electrons flow-

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* Decimal classification: R282.12. Original manuscript received by the Institute, January 20, 1953; revised manuscript received May 7, 1953.

† Bell Telephone Laboratories, Inc., Murray Hill, New Jersey.

[1] W. Shockley, "A unipolar field effect transistor," PROC. I.R.E., vol. 40, pp. 1365–1376; November 1952.

ing through the n-type *channel* from source to drain. For small applied voltages the resistance R_0 of this channel is

$$R_0 = L/2aZ\sigma \qquad (1)$$

where σ is the conductivity of the n-type germanium, L is the length of the channel, $2a$ its thickness and Z its width. As the drain voltage is raised, more current flows in the channel, and the IR drop along the channel makes it more positive the greater the distance from the source. This means that those portions of the gate p-n junction which are most distant from the source will be biased most strongly in the reverse direction and a wedge shaped space-charge barrier will result. This space charge region is shown by the dotted areas of Fig. 1. The current flow is confined to the neutral channel (shown white in the figure). When the drain voltage reaches a critical limiting value, W_0, the channel is completely *pinched off* at its drain end. Virtually no further increases of drain current will result when the drain voltage is raised above W_0, the only effect of this voltage being to change the channel shape at constant current. The value of this *pinch-off voltage* is given by

$$W_0 = \rho_0 a^2/2K = \sigma a^2/2\mu K = qNa^2/2K \qquad (2)$$

where a is the half-thickness of the channel, $\rho_0 = qN$ is the magnitude of the fixed donor charge density in the space-charge region, and K is the dielectric constant. For germanium

$$K = 1.42 \times 10^{-12} \text{ farads/cm.}$$

The use of K in farads/cm rather than $\kappa\epsilon_0$ in farads/m can be justified as indicated in ref. 1, page 1367.

Now suppose that the gate is biased negatively with respect to the source. Under these conditions the magnitude of the IR drop necessary to produce pinch-off will be smaller, since part of this voltage is already supplied by the bias, and saturation will accordingly occur at a lower value of drain voltage and current. Shockley has shown that, if the wedge-shaped channel referred to above narrows sufficiently slowly, then, for drain voltages less than the saturation value of V_D, the current (per unit length in the Z direction) is given by[2]

$$I_D = (1/L)\left[J(V_D - V_G) - J(V_S - V_G)\right] \qquad (3)$$

where the function J is given by

$$\begin{aligned} J(x) &= 2\sigma a x\left[1 - (2/3)(2xK/\rho a^2)^{1/2}\right] \\ &= g_0 x\left[1 - (2/3)(x/W_0)^{1/2}\right]. \end{aligned} \qquad (4)$$

Above the saturation value of V_D the current is substantially constant.

It will be convenient to introduce certain natural parameters defined as follows:

$$\begin{aligned} W_0 &= \rho_0 a^2/2K \\ E_0 &= \rho_0 a/K = 2W_0/a \\ g_0 &= 2\rho_0\mu_0 a = 2\sigma a \\ I_0 &= g_0 E_0 = 2\rho_0{}^2\mu a^2/K \\ \tau_0 &= a/\mu_0 E_0 = K/\mu_0\rho_0 = K/\sigma. \end{aligned}$$

In Fig. 2 we have shown the drain voltage-current characteristics as obtained from (3) and (4). It can be seen from the form of (3) that different values of V_G translate the curve in both the I_D and V_D directions. As a consequence of this, the curve joining the saturation points of the characteristics (shown dotted) is simply the $V_G = 0$ curve rotated through 180°.

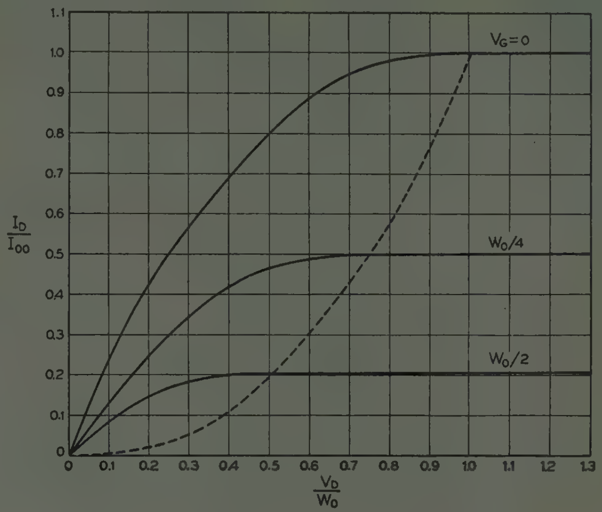

Fig. 2—Normalized drain characteristics with source at zero potential. The dashed curve shows the locus of the pinch-off points (see text).

Comparisons of these theoretical curves with those experimentally determined will be made in Part II. The gate currents, in general, will be the reverse saturation current of the gate p-n junction and can be neglected.

We are now in a position to calculate the transconductance of the device. From (3) and (4) we obtain

$$\begin{aligned} g_m &= \left.\frac{\partial I_D}{\partial V_G}\right|_{V_D = \text{cnst}} \\ &= (2\sigma a/LW_0{}^{1/2})\left[(V_D - V_G)^{1/2} - (V_S - V_G)^{1/2}\right]. \end{aligned} \qquad (5)$$

Now if in particular we take $V_S = 0$ and $(V_D - V_G) = W_0$ corresponding to grounded source and operation in the pinch-off region, this reduces to

$$\begin{aligned} g_{mG} &= (2\sigma a/L)\left[1 - (-V_G/W_0)^{1/2}\right] \\ &= g_{m0}\left[1 - (-V_G/W_0)^{1/2}\right] \end{aligned} \qquad (6)$$

where we have introduced

$$g_{m0} = \frac{2\sigma a}{L} \qquad (7)$$

the maximum transconductance. This quantity can be obtained from the nomograph of Fig. 4.

The saturation drain current I_{DG} which flows for any gate voltage can be obtained from (3) and (4) by setting $V_S = 0$; $(V_D - V_G) = W_0$. This yields

$$I_{DG} = (g_{m0}W_0/3)\left[1 + (V_G/W_0)(3 - 2\sqrt{-V_G/W_0})\right]. \qquad (8)$$

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[2] W. Shockley has discussed the case of a p-type channel. For the n-type channel appropriate changes of sign must be made.

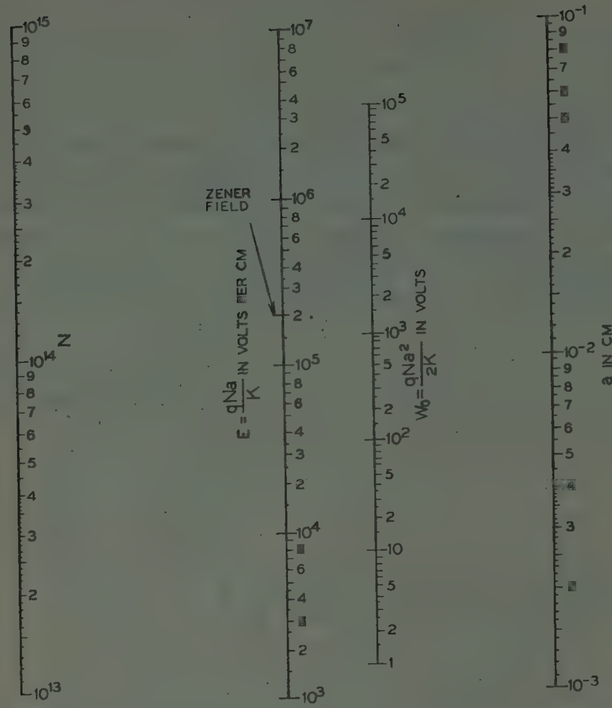

Fig. 3—Pinch-off nomograph.

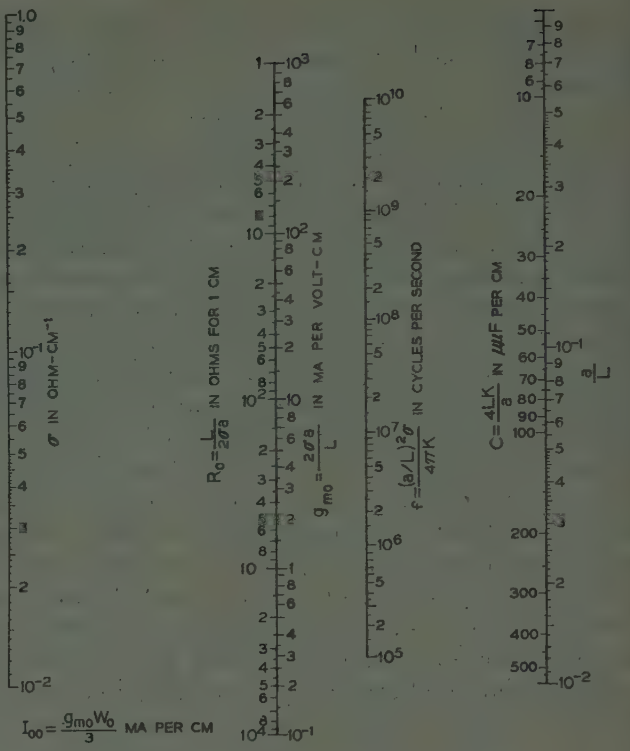
σ IN OHM-CM^-1
1.0
10^-1
10^-2
I_00 = g_mo W_0/3 MA PER CM
R_0 = 1/2σa IN OHMS FOR 1 CM
g_mo = 2σa/L IN MA PER VOLT-CM
10^3
10^2
10
1
10^-1
f = (a/L)^2 σ/4πK IN CYCLES PER SECOND
10^10
10^9
10^8
10^7
10^6
10^5
C = 4LK/a IN μμF PER CM
a/L
10^-1
10^-2

Fig. 4—Field-effect transistor nomograph.

We see from this that the current is completely cutoff for a gate voltage of $V_G = -W_0$. Maximum current is

$$I_{D0} = g_{m0} W_0 / 3 \tag{9}$$

obtained for zero bias on the gate.

The frequency response of the device can be estimated by the following simple argument. In order to change the gate voltage the capacity of its *p-n* junctions must be charged through the resistance of the channel. This process has an associated time constant which limits the frequency response. Let us assume a wedge-shaped channel, completely pinched off at the drain end and completely open at the source end (that is $V_G = V_S = 0$). The capacity for unit length in the Z direction is approximately

$$C \simeq 4KL/a. \tag{10}$$

The factor 4 arises because the average width of the space-charge region is approximately $a/2$ and because there are two such regions, one on either side. This capacity on the average charges through half the resistance of the channel, i.e.

$$R = L/2a\sigma. \tag{11}$$

We would accordingly expect a limiting frequency f given by

$$f = 1/2\pi RC = a^2\sigma/4\pi L^2 K. \tag{12}$$

These relationships are also shown on the nomograph of Fig. 4.

3. *Modifications of the Ideal Theory*

The theory presented in Section 2 deals with an ideal structure. In practice we shall find it necessary to modify the theory somewhat to take account of special experimental conditions. In particular we shall discuss the following effects:

A. Series resistance of semiconducting paths at the source and/or drain contacts.

B. Depletion of thermally generated carriers in units having conductivity close to intrinsic.

C. Negative resistance effects due to hole current flow into the gate.

D. Temperature effects.

A. Series Resistance. In Section 2 we have considered that source and drain connect directly onto the channel between the gates. It has been found necessary in the fabrication of these units, however, to allow a small bridge of semiconductor between the actual contact and the gate, an example of which is shown in Fig. 5. This means that a series resistance appears between the electrodes to which voltages are applied and the working part of the structure. It is possible to take account of these resistances by simple circuit theory.

For the calculation of g_m one must recognize two differences. Firstly, the voltages which must be inserted into (5) to obtain the transconductance of the working part of the structure must be changed as follows:

$$V_G \rightarrow V_G,\ V_S \rightarrow -IR_S,\ V_D \rightarrow (V_D - IR_D) \tag{13}$$

where R_S and R_D are the values of the source and drain resistances. Secondly, the degeneration of the source resistor must be taken into account. If the apparent transconductance is denoted by g_m', we must write

$$g_m' = g_m/(1 + R_S g_m) \tag{14}$$

which takes account of the fact that a fraction of any increase in gate voltage appears as an increased drop across the source resistor. In (14), of course, g_m must be calculated considering the modifications (13).

It is clear from (14) that if $g_m R_S \gg 1$, then the transconductance g_m' is simply given by $1/R_S$. If then we are to obtain the potentially high transconductance of the device, we must keep the source resistance small. The drain resistance does not have as serious consequences, the chief disadvantages being, (a) the necessity of having a higher supply voltage, and (b) $I_D^2 R_D$ heat which must be dissipated.

The source resistance also alters the frequency response. One must add R_S to the R of (12) in obtaining the limiting frequency.

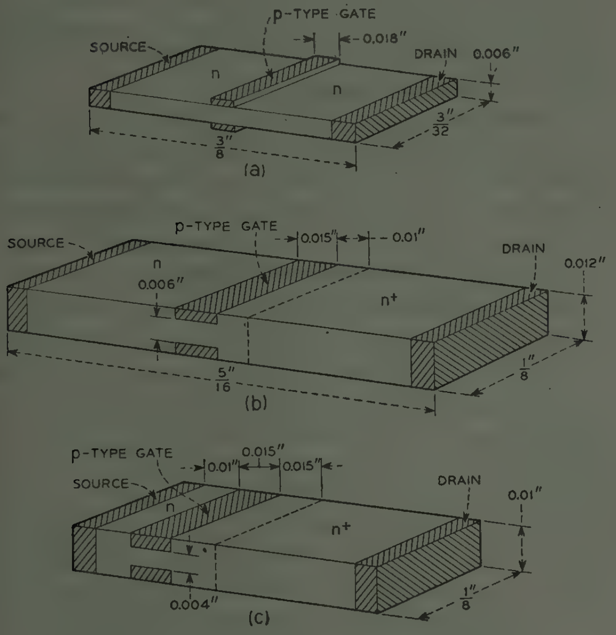

Fig. 5—Schematic diagrams of experimental units.

B. Carrier Depletion. Suppose, as in Fig. 5(b), that at the drain end the germanium is made strongly n-type (designated n^+). Such a n-n^+ junction will not act as a source of holes. Now the direction of current flow is such that electrons tend to flow into the drain, and holes into the source. These holes must largely arise by generation processes within the volume of the semi-conductor or at the free surfaces, and these holes will tend to be depleted. In order to maintain electrical neutrality within the body of the crystal an equal density of electrons must also be depleted. In the limit of strong applied fields, such that carriers are swept out in a small fraction of a mean lifetime, virtually all of the holes will be swept out and the only remaining carriers will be electrons equal in density to the excess donor density. In material which is close to intrinsic conductivity this means the apparent conductivity will go down. In fact

$$\sigma = q\mu_n N$$

instead of

$$\sigma_i = q\mu_n n_i(1 + 1/b)$$

so that the conductivity has changed in the ratio

$$\sigma/\sigma_i = (Nb/n_i)/(1 + b) \approx 2N/3n_i. \quad (15)$$

C. Negative Gate Resistance. It is possible, in units with no n-n^+ junction where an appreciable fraction of the current is carried by holes, to obtain a negative gate resistance. This effect arises as follows: Suppose that the gate is made more positive so that it opens somewhat and allows additional electron current to flow in the channel. This additional electron current is accompanied by an increased hole current on the drain side of the gate, since electrical neutrality must be maintained. If the unit is operating in the pinched-off region, however, these holes will not flow through the channel but will instead flow to the gate. This is the case because, while the electric field in the pinched-off space-charge region near the drain is directed away from the axis of the channel and is thus focussing for electrons, it tends to pull holes into the gate. From this argument we see that a positive change of gate voltage results in an increased flow of holes to the gate and thus in a negative change in the current flowing into the gate. Hence the gate terminal presents a negative resistance.

The holes referred to in the argument above may arise from three sources: (a) They may be thermally generated in the body of the semiconductor; (b) they may be generated at low life time areas on the surface; or (c) they may be injected at the contacts. It might be desirable to make a device in which the drain contact was intentionally made hole-injecting (say by the use of a p-n junction) and in which the negative resistance effects would thus be greater.

D. Temperature Effects. As can be seen from the nomographs, the unipolar field effect transistor is a relatively high power device. Furthermore, most of the power dissipation takes place in the space-charge region of the drain. Therefore the removal of heat will be a major problem. These devices should probably not prove as temperature sensitive as the n-p-n transistor. The chief effect of heating is to increase the gate saturation current by increasing the number of thermally generated hole-electron pairs. It is true that the mobility also varies as $T^{-3/2}$ but this variation is fairly slow and only serves to vary σ which enters the theory linearly. In Part II experiments are described which were performed at liquid air temperatures. Here, of course, the mobility has increased by a factor of about 8 times and major differences arise.

4. *Design Considerations*

The theory of the preceding sections has been summarized in the form of nomographs which are given in Fig. 3 and Fig. 4. In Fig. 3 we have given a nomograph for the calculation of the pinch-off voltage W_0. If a straight line is drawn between the value of N on the left scale and the value of a on the right scale, it intersects the center scales at W_0 and at E, the maximum value of the electric field in the space charge region. Fig. 4 shows simultaneously the consistent field-effect parameters, a/L, σ, g_{m0}, R_0, f and C. A straight line drawn across the chart intersects the various scales in a set of values which are consistent with a given design. The one remaining parameter of interest, the current, is easily obtained from (9).

It is clear from an inspection of the nomographs that the choice of material and dimensions is not completely arbitrary. It is desirable to have the pinch-off voltage reasonably low both in order to avoid heating and to avoid nonlinear mobility effects which set in at high electric fields and which would limit the transconductance. A low pinch-off voltage implies small spacing and/or small donor concentration. Small donor concentrations are undesirable however because of depletion effects and because both f and g_m are proportional to σ. The best compromise therefore is to make a and L/a as small as is mechanically feasible and then to choose N in order to make W_0 reasonably small.

A typical design example is shown below

$$N = 10^{14}$$

$$\sigma = .059 \text{ mho/cm}$$

$$a = 3 \times 10^{-3} \text{ cm}$$

$$W_0 = 50 \text{ volts}$$

$$a/L = .2$$

$$g_{m0} = 24 \text{ ma/volt cm}$$

$$f = 140\ Mc/S$$

PART II, EXPERIMENTAL RESULTS

1. *Introduction*

This part contains a description of the experimental work that was done in order to test the theory of unipolar "field-effect" transistors. As discussed in Part I, section 4, a suitable choice of resistivity for the channel material would be about 20 ohm-cm. However, in order to have a pinch-off voltage of about 10 volts with such material, the channel width would have to be about 10^{-3} cm and such a structure would be difficult to make. If higher resistivity material, say 50 ohm-cm, were used, the channel width for 10 volts pinch-off would be of the order of 4×10^{-3} cm and preparation would be much simpler. However, with such material, the effects of depletion of carriers would be appreciable and there would be a consequent loss of transconductance and frequency response. It was decided that, for the purposes of verifying the theory, performance characteristics could be sacrificed to ease of preparation and 47 ohm-cm n-type germanium was chosen.

Units were made with the three geometries shown in Fig. 5(a), (b) and (c). For clarity, the diagrams show the gates as being only on the top and bottom surfaces of the germanium blank. In practice, the gates extended over the sides of the blank to make a continuous loop. This was done in order to insure that none of the channel current flowed along the surfaces of the germanium. Pearson and Read had found that free surfaces in the channel region resulted in instability in the characteristics of their specimens. No special means were provided for removing heat generated in the specimens. This would have required attaching cooling fins or a heat sink of some form to one of the three contacts. It was decided that the added complication was not justified in a preliminary survey of properties of the device.

In all, six specimens were made and tested. This part will include a discussion of three of these which exhibited all the salient features of the others.

2. *Specimen PF 66 #2*

The dimensions of the specimen are shown in Fig. 5(a). This was the first specimen to be made and was designed for ease of fabrication. Ohmic contacts for the source and drain were made at the ends of the germanium blank and the gates were situated approximately half way between source and drain. The theoretical pinch-off voltage, as obtained from the nomograph of Fig. 3, for a channel half-width of 0.003″ in 47 ohm-cm n-type material is 32 volts.

With the specimen at 25°C and at 0°C, the static characteristics were measured and are reproduced in Figs. 6 and 7. Fig. 6 shows the $I_D - V_D$ curves with V_G as parameter. The curves have the same form as the theoretical curves of Fig. 2. The experimental values of pinch-off voltage and transconductance at 25°C can be compared with the theoretical values as follows:

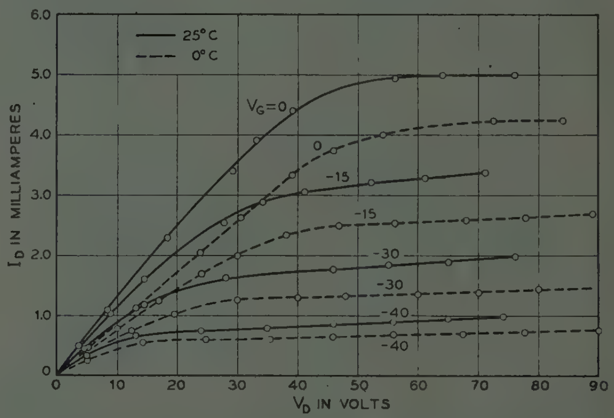

Fig. 6—Drain characteristics of unit PF66 #2. The solid curves are for 25°C and the dashed curves for 0°C.

It is first necessary to know the values of the source and drain resistances, R_S and R_D. The method of preparation of the specimen did not permit of accurate measurements of the source-to-gate and gate-to-drain distances and it is therefore impossible to calculate R_S and R_D directly. However, if V_G is zero and V_D small compared to W_0, the resistance between source and drain will be essentially the sum of R_S, R_D and R_0. Hence this sum can be determined from the initial gradient of the $I_D - V_D$ curve for $V_G = 0$. From this initial slope

$$R_S + R_D + R_0 = 7.8\ K\Omega.$$

From the nomograph of Fig. 4

$$R_0 = 400\ \Omega$$

$$\therefore\quad R_S + R_D = 7.4\ K\Omega.$$

Since the gate was half way between source and drain it is reasonable to take

$$R_S = R_D = 3.7\ K\Omega.$$

Now, for $V_G=0$, I_D saturated at $5mA$ with a drain voltage of 55 volts. The IR drop in R_D was then 18.5 volts. Hence

$$W_0 = 55 - 18.5 = 36.5 \text{ volts.}$$

This value is in fair agreement with the theoretical value of 32 volts. The transconductance of the channel for $V_G=0$ can be determined from (5)

$$g_m = \frac{2\sigma a}{L}\left[\frac{(V_D - V_G)^{1/2} - (V_S - V_G)^{1/2}}{(W_0)^{1/2}}\right] \qquad (5)$$

by putting $2\sigma a/L = g_{m0}$, $V_G=0$ and $V_D - V_G = W_0$ giving

$$g_{mS} = g_{m0}\left[1 - \left(\frac{V_S}{W_0}\right)^{1/2}\right].$$

In this case g_{m0} as determined from Fig. 4 is $1.67\ mA/V$,

$$W_0 = 36.5V; \quad V_S = R_S(I_D - I_G) = 3.7(5 - 1.5) = 13V$$

$$\therefore \quad g_{mS} = 0.67\ mA/V.$$

The effective transconductance g_m', for $V_G=0$, is then obtained from (14)

$$g_m' = \frac{g_m}{1 + R_S g_m} = 0.19\ mA/V.$$

From Fig. 6 the average transconductance over a range of V_G from 0 to -15 volts is found to be 0.12 mA/V. This value is smaller than the calculated value for $V_G=0$ as would be expected since (6) shows that the transconductance decreases as V_G becomes more negative. A more useful comparison could be made with the calculated value of g_m' for $V_G=-7.5$ volts. For these conditions (5) and (14) give $g_m'=0.16\ mA/V$. The agreement between this and observed value of 0.12 mA/V is within the accuracy of the calculations.

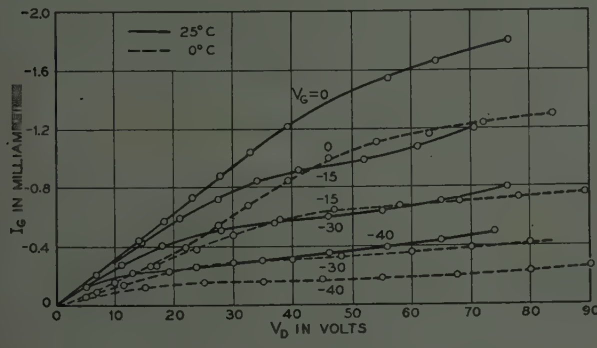

Fig. 7—Gate current versus drain voltage for specimen PF66 #2.

Fig. 7 shows the $I_G - V_D$ curves with V_G as parameter. As discussed in Part I, section 3, the gate current consists of holes which could arise in three ways: thermal generation in the germanium between gate and drain, generation at a surface of the germanium, and injection at the drain. For this specimen, the gate currents obtained from the first two causes would be of the order of $100\mu A$ and would be very temperature dependent. Fig. 7 shows that the gate current was as high as $2mA$ and that a temperature change of 25°C produced only 30 per cent change in I_G. This suggests that the gate current arose predominantly from hole injection at the drain. If this were the case, then the gate current and drain current at any particular values of V_D and V_G should be related by the equation

$$\frac{I_D}{I_G} = \frac{n_0\mu_n E + p_0\mu_p E}{p_0\mu_p E} = \frac{n_0\mu_n + p_0\mu_p}{p_0\mu_p}$$

where n_0 and p_0 are the equilibrium densities of electrons and holes, μ_n and μ_p their mobilities, and E the electric field at the drain contact.

For this material

$$n_0 = 3 \times 10^{13}/\text{cm}^3$$

$$p_0 = 2 \times 10^{13}/\text{cm}^3$$

$$\mu_n = 3600 \text{ cm}^2/\text{volt-sec}$$

$$\mu_p = 1700 \text{ cm}^2/\text{volt-sec}$$

$$\therefore \quad \frac{I_D}{I_G} = 4.08$$

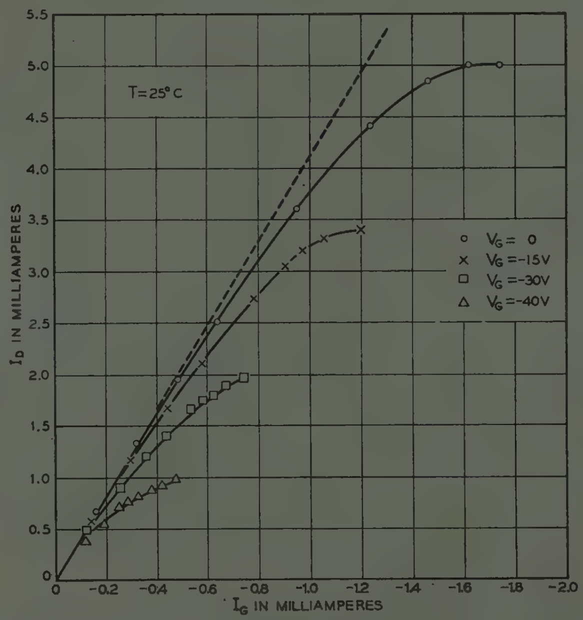

Fig. 7(a)—Drain current versus gate current with gate voltage as parameter for specimen PF66 #2. The theoretical relationship is shown in the dashed curve.

Fig. 7(a) gives a plot of I_D vs. I_G with V_G as parameter, the progression towards higher currents on each curve corresponding to increasing values of V_D. The broken line is the theoretical relationship. It is seen that initially the experimental curves follow the theoretical one but that at higher currents the gate current increases more rapidly. This deviation of gate current is of the order of magnitude that would be obtained from holes generated in the body and at the surfaces of the germanium. It may be concluded that the gate current arose from all three causes but that injection at the drain was the predominant effect.

In Fig. 8 the data of Fig. 7 has been replotted to give the gate characteristic. The figure shows that the gate exhibits negative resistance, as predicted by the theory of Part I, section 3, and that its magnitude varies from 30 KΩ at $V_D=60$ volts to very large values at $V_D=10$

volts. It should therefore be possible to make the unit oscillate by putting an $L-C$ resonant-circuit in series with the gate. This was done with the specimen operating at $V_D=60$ volts and $V_G=0$ and oscillations were produced up to a frequency of 100 kc. It was found that the oscillations were killed by shunting the resonant circuit with a resistance of less than about 30 KΩ which is the magnitude of the negative resistance at these operating conditions.

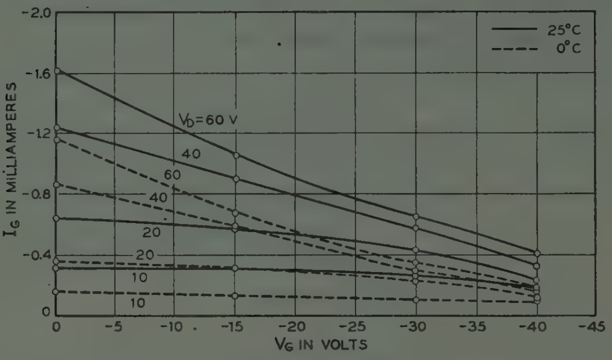

Fig. 8—Gate characteristic for specimen PF66 #2. The solid lines refer to 25°C and the dashed lines to 0°C.

The measurements on this specimen showed that this form of structure would result in an amplifying transistor in agreement with the theory and that emission of holes from the drain caused the gate to have a negative resistance. If a unit were required as a negative-resistance oscillator, then it would be desirable to enhance the emission of holes from the drain. This could be achieved by using a *p-n* junction as the drain contact. However, for an amplifying transistor, this negative gate-resistance and the high values of gate current associated with it would both be undesirable. For this application, emission of holes from the drain should be reduced to a minimum. This could be achieved by using an $n-n^+$ junction for the drain contact. The next specimen to be described, *PF* 70 #7, was made this way.

3. *Specimen PF 70 #7*

The dimensions of the specimen are shown in Fig. 5(b). The resistivity of the *n* material was 47 ohm-cms, and the resistivity of the n^+ material was less than 1 ohm-cm. To reduce the value of R_D the gate was placed as close as possible to the drain (approximately .01″).

As discussed in Part I, section 3, virtually all the holes will be swept out of the structure if the transit time for holes is small compared to the lifetime. For this specimen the transit time for holes across the distance l from drain to source will be approximately

$$\frac{l^2}{\mu_p V_D} = \frac{\left(\frac{5}{32}\right)^2}{1700\ V_D} = \frac{15}{V_D} \times 10^{-6} \text{ secs.}$$

Therefore, if $V_D=1$ volt, the transit time will be of the order of 15μ secs which was small compared to the lifetime of the material. It is thus reasonable to assume that, at all normal operating conditions, the material will be swept free of holes. Conductivity will then be maintained by the remaining electrons which are just those required to neutralize the excess charge-density of the excess donors, in this case 10^{13}/cc. The effective conductivity will therefore be

$$10^{13} \times 1.6 \times 10^{-19} \times 3600 = 5.8 \times 10^{-3} \text{ mho/cm.}$$

The estimated pinch-off voltage from Fig. 3 is 32 volts.

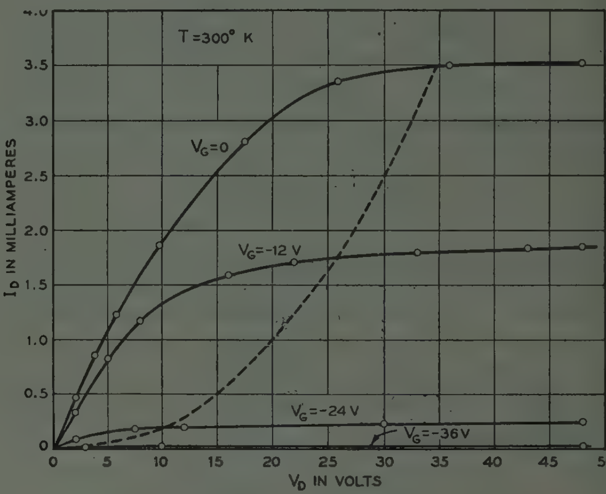

Fig. 9—Drain characteristic at 300°K for PF20 #7. The dashed line is the same shape as the $V_G=0$ line but is rotated through 180°. Note this curve is approximately the locus of pinch-off points.

With the specimen at 300°K and at 77°K, the static characteristics were measured and are reproduced in Figs. 9–12. Fig. 9 shows the I_D-V_D curve with V_G as parameter and the specimen at 300°K. The theoretical values of pinch-off voltage and transconductance can be estimated as for the previous specimen. In this case, the magnitude of R_D is sufficiently small so that the *IR* drop in it can be neglected and, hence, the value of V_D at saturation is equal to W_0. The observed value of W_0 is then about 35 volts and the theoretical value 32 volts. The observed value of g_m' over a range of V_G from 0 to -12 volts is 0.14 mA/V while the calculated value at $V_G=-6$ volts is 0.19 mA/V. Hence, the theory and experiment are in reasonable agreement.

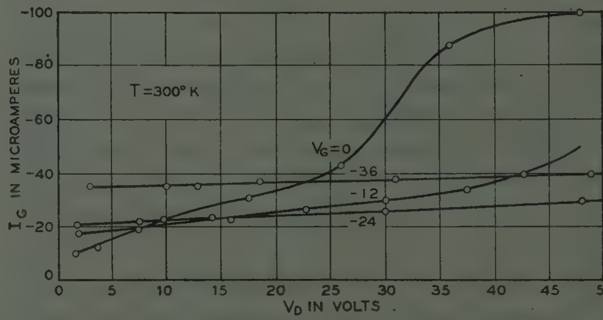

Fig. 10—Gate current versus drain voltage for specimen PF70 #7 at 300°K.

Fig. 10 shows the plot of I_G-V_D with V_G as parameter. It is seen that the gate current is now less than 100 μA. Injection of holes at the drain had therefore been greatly reduced or eliminated. Also I_G increased most rapidly with V_D under conditions where V_D was large and V_G small. Since these are the conditions under which the power dissipated in the unit was greatest and hence the temperature highest, the evidence suggests that the source of current was thermal generation of holes. The data of Fig. 10 have been replotted in Fig. 11 to give the gate characteristic. When the power dissipation was greatest the gate exhibited negative resistance. This is in qualitative agreement with theory since under these conditions the temperature, and therefore rate of generation of holes, is highest and therefore the proportion of current carried by holes is greatest.

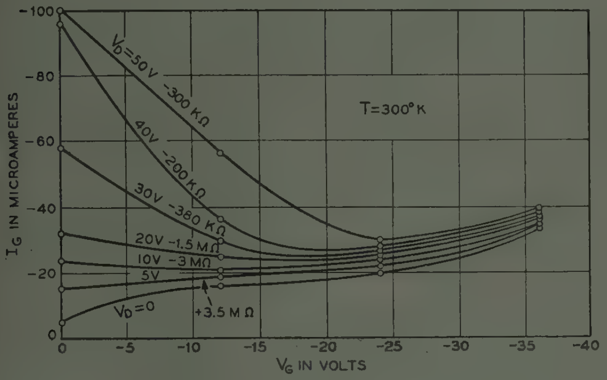

Fig. 11—Gate characteristic of specimen PF70 #7 at 300°K.

As shown in Part I, section 3, negative resistance can occur under such conditions. When the power dissipated and hence the temperature and rate of hole generation was lowest, the gate exhibited positive resistance. The negative resistance of the gate was always greater than 200 $K\Omega$ and therefore the specimen should be stable, provided the impedance in the external circuit to the gate is less than this value. The specimen could therefore be conveniently used as an amplifier. Furthermore, if means were provided to take away the heat generated in the unit, as by attaching cooling fins, negative resistance of the gate could be further increased.

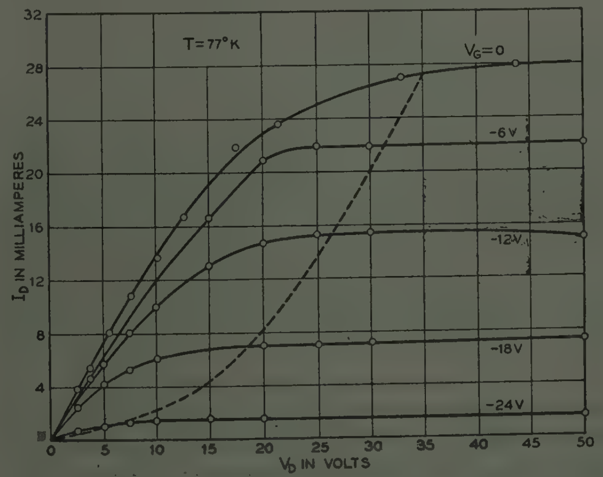

Fig. 12—Drain characteristic of specimen PF70 #7 at 77°K. Note reversed $V_G=0$ curve shown dashed.

Fig. 12 shows the I_D-V_D curves with the specimen at 77°K. The observed value of W_0 is 35 volts compared to a theoretical value of 32 volts. The observed value of g_m' over a range of V_G from 0 to -6 volts is $1 mA/V$ while the calculated value at $V_G=-3$ volts is 0.95 mA/V. The gate current was found to be always less than 1 μA. At this temperature the thermal generation of holes should have been very small. Therefore, so must the injection of holes at the drain.

4. *Specimen PF 73 #10*

The dimensions of the specimen are shown in Fig. 5(c). The source-to-gate distance was made as small as possible in order to decrease the value of R_S and hence increase the effective transconductance. The width of the channel was decreased in an attempt to achieve a lower value of pinch-off voltage.

With the specimen at 300°K and at 77°K, the static characteristics were measured and are reproduced in Figs. 13 to 16. Fig. 13 shows that I_D-V_D curve with V_G as parameter and the specimen at 300°K. The resistance as calculated from the initial slope of the $V_G=0$ curves is 2.2 $K\Omega$ while the channel resistance, R_0, as determined from Fig. 4 is also 2.2 $K\Omega$. It may therefore be concluded that effectively all the resistance occurred in the channel and that both R_S and R_D may be neglected. The calculated pinch-off voltage is 14 volts while the observed value was about 12 volts. The observed transconductance over a range of V_G from 0 to -3 volts was 0.32 mA/V while the calculated value at $V_G=-1.5$ volts is 0.30 mA/V. Figs. 14 and 15 show the I_G characteristics which are similar in form to those for specimen *PF* 70 #7. The frequency response of the specimen was estimated by connecting it as a tuned-drain feedback oscillator and determining the maximum frequency at which it was possible to excite oscillation. The specimen was made to oscillate up to frequencies of 4 mc/s. This is in reasonable agreement with the theoretical maximum, as determined from Fig. 4, of 6 mc/s.

Fig. 16 shows the I_D-V_D curves for the specimen at 77°K. The observed pinch-off voltage was 12 volts compared to the theoretical value of 14 volts. The transconductance over a range of V_G from 0 to -3 volts was 1.15 mA/V as compared to the calculated value at $V_G=-1.5$ volts of 2.8 mA/V. The observed value of transconductance was therefore much less than the calculated value and the discrepancy is too large to be attributed to experimental error. A possible explanation would be that the field in the channel is so high that the mobility is no longer independent of field. Ryder has shown[8] that, above a certain critical value of field, the mobility is proportional to $1/E^{1/2}$ and that, for the *n*-type germanium at 77°K, the critical value is about 140 volts/cm. In this specimen, virtually all the potential drop of 12 volts would occur in the channel region. The average field in the channel was therefore $12\times1000/15\times2.54=310$ volts/cm which is greater than the critical field. Since the field at the drain end of the channel is greater than the average field, at least part of the channel had a field greater than critical.

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[8] E. W. Ryder, to be submitted to Phys. Rev. See also: W. Shockley, "Hot electrons in germanium and Ohm's law," *Bell System Tech. Jour.*, vol. 30, pp. 990–1034; October, 1951.

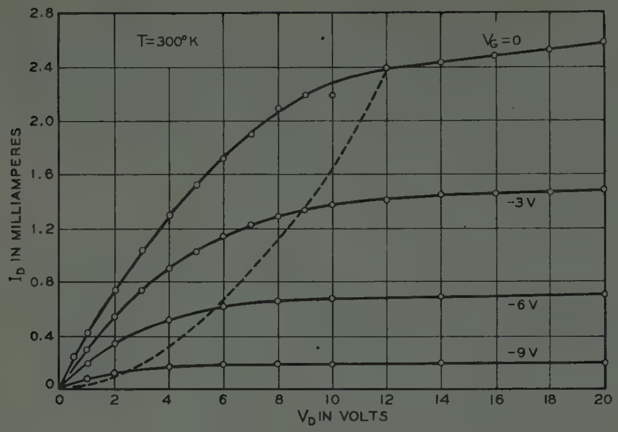

Fig. 13—Drain characteristic of specimen PF73 #10 at 300°K. Note locus of pinch-off points.

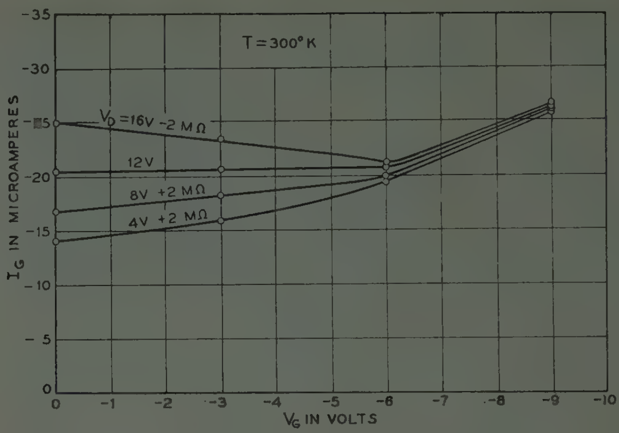

Fig. 15—Gate characteristic for specimen PF73 #10 at 300°K.

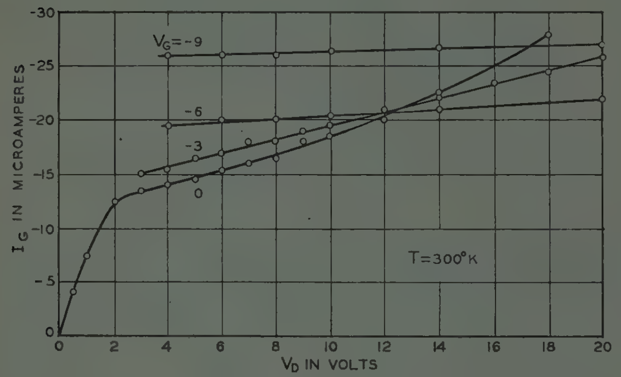

Fig. 14—Gate current versus drain voltage for specimen PF73 #10 at 300°K.

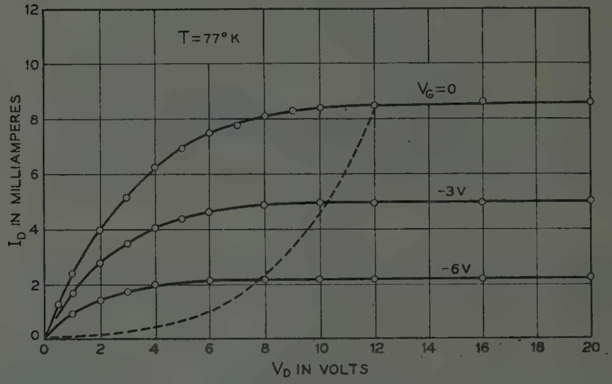

Fig. 16—Drain characteristic for PF73 #10 at 77°K.

Hence the effective conductivity and therefore the transconductance would be less than that calculated on the assumption of constant mobility. At 300°K the value of the critical field is about 900 volts/cm which is much greater than the maximum field present in the specimen. Hence at 300°K the theoretical and experimental values of transconductance should agree. The variation of mobility therefore seems capable of explaining the observed results.

5. *Stability and Noise*

The *D-C* characteristics of all the specimens were found to be stable with time. The characteristics were repeatable to a high accuracy. The specimens could be subjected to fairly high temperatures without changing the characteristics at room temperature. For example, the power dissipation in one specimen was on one occasion so great that the soldered gate connection melted. New leads were soldered on, the specimen re-etched and the characteristics found to be unchanged.

The noise properties of specimen *PF* 73 #10 were examined by H. C. Montgomery.[4] With the specimen at 300°K, $V_D=17.5$ volts, $V_G=-3$ volts and a power gain of 19.6 *db*, the noise figure at 1000 c/S was 68 *db*. This is large compared to the noise figure for an average *n-p-n* of about 10-15 *db*. At 77°K, with $V_D=12.5$ volts, $V_G=-4.7$ volts and a power gain of 30.2 *db*, the noise figure was about 35 *db*. By varying the bias conditions a minimum noise figure of 25 *db* was measured. The most significant change in the characteristics of the specimen in going from 77°K to 300°K is the increase in gate current. It may then be that the difference in noise figure can be attributed to the gate current. If this is the case, suppression of gate current at 300°K would lead to a unit having a noise figure comparable to an *n-p-n* transistor.

Part III, Conclusions

The experimental results of Part II are substantially in agreement with the extended theory. Transistors have been made which at room temperature had transconductances as high as 0.3 ma/volt, and whose characteristics were very stable. A flat frequency response has been obtained up to about 3 mc/sec. In addition a new feature has been observed, namely negative gate resistance. The gate resistance can be made either positive or negative depending on the nature of the drain contact.

There is reason for confidence, based on the agreement with theory which has already been obtained, that the theory can be used to design transistors for higher transconductances and frequency response. Such a design was discussed in Part I sec. 4 where it was shown that using 17 ohm-cm germanium a transconductance of 24 ma/volt and an upper frequency limit of 140 mc/sec could be obtained. This does not necessarily represent a limiting design but should be well within the scope of present fabrication techniques.

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[4] H. C. Montgomery, personal communication.

The chief virtue of this new form of transistor is its high frequency response. To obtain similar frequency response with point contact or junction transistors would require structures with much smaller critical dimensions. This is a consequence of the unipolar nature of the device, that is, that essentially only the majority carriers are involved in its operation. Another important feature of field-effect transistors is their high input and output impedance together with high transconductance. In this respect they resemble pentodes. They are more efficient than pentodes, however, in that they do not require cathode heating power.

While, as stated above, it is possible to achieve high transconductance and frequency response, this can only be done at the expense of high current and power dissipation. Since, in addition, these units have a relatively high noise figure, it appears that they will not be suitable for low power applications. Units have been made of silicon[5] which have much lower noise figure but these have lower transconductance and frequency response.

Acknowledgments

The authors wish to acknowledge the help of P. W. Foy and W. Wiegmann in the fabrication of the units and of W. W. Bradley who supplied the germanium. They are also indebted to W. Shockley for advice and encouragement.

Symbols

a—half thickness of channel
b—μ_n/μ_p
C—capacitance
E—maximum value of electric field in space-charge region
E_0—$2W_0/a$
f—frequency
g—conductance
g_0—$2\rho_0\mu a$
g_m—transconductance
I_D—current into drain
I_G—current into gate
I_S—current into source
I_0—g_0E_0
K—dielectric constant
L—length of channel
N—excess impurity concentration
N_d, N_a—density of donors, acceptors
q—electronic charge
R—resistance of channel
R_D—resistance of drain contact
R_S—resistance of source contact
R_0—resistance of channel at zero bias
τ_0—K/σ
V_D—potential on drain
V_G—potential on gate
V_S—potential on source
W—potential between channel and gate
W_0—value of W for pinch-off
Z—width of channel
ρ_0—fixed charge density
σ—conductivity
σ_i—intrinsic conductivity
μ_n, μ_p—mobility of electrons, holes

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[5] G. L. Pearson, *Phys. Rev.*, vol. 90, p. 336; 1953.

A Subjective Study of Color Synchronization Performance*

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Summary—The NTSC color-synchronization signal is reviewed to determine its operational quality. A subjective testing system is then described which dissects various receiver functions such that the thermal-noise impairment in each function may be evaluated. Results of these subjective tests show that color-synchronization performance is completely adequate under conditions of thermal-noise interference. The tests also show that the limiting factors of performance are primarily those functions which are common to monochrome transmission.

THE HIGH QUALITY of the NTSC color television[1,2] system is primarily obtained by the transmission of a greater amount of useful information through the conventional television channel. This increase has been accomplished by utilization of the vacant frequency spectrum of monochrome television and by quadrature modulation of the color information upon a 3.898125 mc subcarrier. Realization of the extra information at the receiver, however, requires the presence of an unmodulated reference signal of the same frequency and of known phase relative to one of the modulation components of the subcarrier. Thus, an extra synchronizing function is necessary in the system to provide the information required to generate the color-reference signal.

In choosing the form of the color-synchronizing signal, it is obvious that several requirements must be met. First, the signal must contain sufficient energy so that adequate-noise performance may be obtained at noise levels so high that the picture is no longer useful.

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* Decimal classification: R583.13. Original manuscript received by the Institute, September 5, 1952; revised manuscript received April 13, 1953.

† Project Engineer, Philco Corporation, Philadelphia, Pa.

[1] B. D. Loughlin, "Recent improvements in band-shared simultaneous color television," PROC. I.R.E.; October, 1951.

[2] C. J. Hirsh, W. F. Bailey, and B. D. Loughlin, "Principles of NTSC compatible color television," *Electronics;* February, 1952.

Quantitatively, this energy level depends upon the required accuracy of the signal, which in the NTSC system is plus or minus 10 to 20°. The signal also must be of such a type that the phase relative to the color-carrier components is maintained under conditions of receiver tuning or drift. Since the phase characteristics of receivers are generally nonlinear in the vicinity of the chroma carrier, compensation for tuning or drift is most simply carried out by using a color-synchronization signal that has the same frequency as the color carrier so that the relative phase shift will remain constant. This argument may be extended to multipath considerations, as well, since small echoes cause the phase of the sync and chroma information to be shifted by the same amount when the frequency of each is identical, so that the relative phase is unchanged. In addition the signal should be such that compatibility is maintained in existing monochrome receivers.

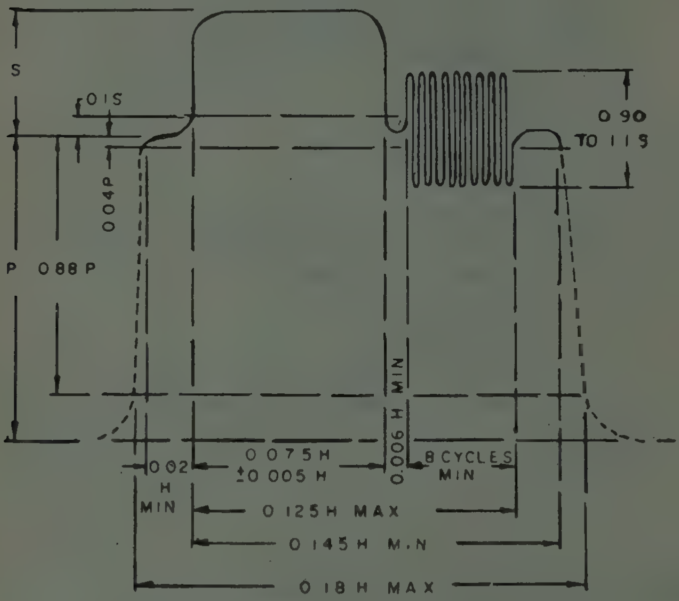

Fig. 1—NTSC compatible color television.

The NTSC specification for the color-sync signal was that adopted by NTSC Panel 14 on May 20, 1952 as shown in Fig. 1 in which a nine-cycle burst of color carrier is inserted upon the back porch of the horizontal-synchronizing signal. At the receiver the burst may be used to create the CW color-reference signal by means of narrow-band filters, locked oscillators, or APC circuits. Since the burst is synchronous in frequency with the color carrier, changes in tuning of the receiver or drift have no effect. This is also true of small echoes as explained previously. It is the purpose of this paper to show the adequacy of the burst method of color synchronization under noise conditions.

Since the merit of a function depends upon how well the intended purpose is accomplished, evaluation of the merit should be made in terms of that purpose. Application of this logic to synchronization in compatible color television suggests that the adequacy of color-synchronization performance should be measured in terms of the basic terminal of the system which is the eye. This may best be accomplished by subjective tests.

Results of such tests depend upon the equipment and operational details; accordingly, these details should be known. The first factor to be considered is restriction of observer reaction to the factor of interest such as the color synchronization. Obviously, a presentation of a noisy signal on a color receiver would show impairment caused by noise in the picture information and deflection synchronizing as well as color synchronization. Thus, noise should be controllable in each of these channels so that impairment may be isolated to a particular function. With this general technique, impairment may be measured for each function of the color system individually so that the separate impairments may be related to each other to determine both absolute and relative subjective performance to a high order of accuracy.

In order to obtain greater dependability and accuracy in the test equipment, the tests were performed by noise matrixing with the composite color video. Fig. 2 illustrates the arrangement of the test equipment in block diagram form.[3] A composite-color video signal was fed to the input of six combiners whose outputs acted as signal sources for the picture-luminosity information, picture-chroma information, deflection synchronization, frame synchronization, and color synchronization for a color receiver, and for a complete monochrome receiver. Noise was introduced into each of the combiners either independently or in combination in such a manner that the operator conducting the test could vary the effective signal-to-noise ratio in a particular channel or channels at will. This allowed the observer to have the noise varied in one or more channels until the impairment corresponded to a particular comment which described his subjective reaction to the impairment caused by the function or functions in relationship to the signal-to-noise ratio.

The noise generator was a modified IF and RF section of a television receiver in which the antenna lead had been replaced by a dummy load and the whole unit well shielded. Normally this type of system would generate thermal noise with a triangular frequency spectrum. However, by increasing the bandwidth slightly and by proper video peaking, the noise spectrum may be made substantially flat throughout the normal television band. After giving effect to these changes, a check of the frequency spectrum of the noise showed it to be flat within 1.5 db of the average level as illustrated in Fig. 3. The chain which handled the noise also was checked so that the uniform spectrum was actually maintained through the combiners. The video and noise levels were set so that when the attenuators were set at minimum the effective signal-to-noise ratio at the output of the

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[3] The technique presented here was originally developed to evaluate the merits of an earlier color system. Results of those tests were described in a paper by E. M. Creamer and M. I. Burgett, Jr., "Performance of Carrier Synchronizing Circuits for Color Television Signals," presented at the March, 1951 IRE Convention.

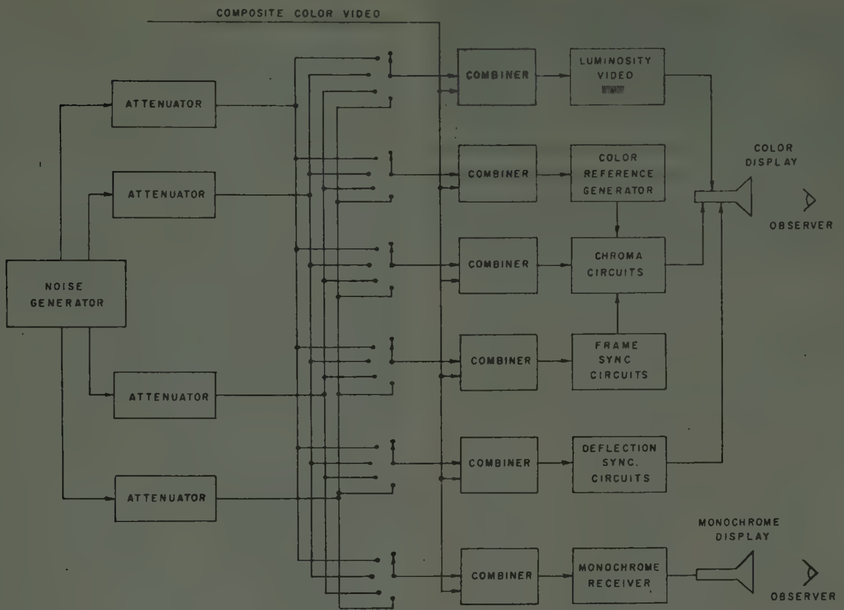
COMPOSITE COLOR VIDEO
ATTENUATOR
ATTENUATOR
NOISE GENERATOR
ATTENUATOR
ATTENUATOR
COMBINER
COMBINER
COMBINER
COMBINER
COMBINER
COMBINER
LUMINOSITY VIDEO
COLOR REFERENCE GENERATOR
CHROMA CIRCUITS
FRAME SYNC CIRCUITS
DEFLECTION SYNC. CIRCUITS
MONOCHROME RECEIVER
COLOR DISPLAY
OBSERVER
MONOCHROME DISPLAY
OBSERVER

Fig. 2—Block diagram of noise and signal matrixing.

combiners was −2 db. Thus, variation of the calibrated attenuator setting could vary the signal-to-noise ratio of a channel in steps of 1 db over the range minus 2 to 60 db. As defined in these tests, signal-to-noise ratio is the ratio of the peak video measured at the sync tips and referenced against white to the RMS value of the noise. It should be noted that this definition gives a signal-to-noise ratio which is pessimistic by a voltage factor of approximately four in comparison with the definition frequently used in which both video and noise are measured as peak-voltage ratios.

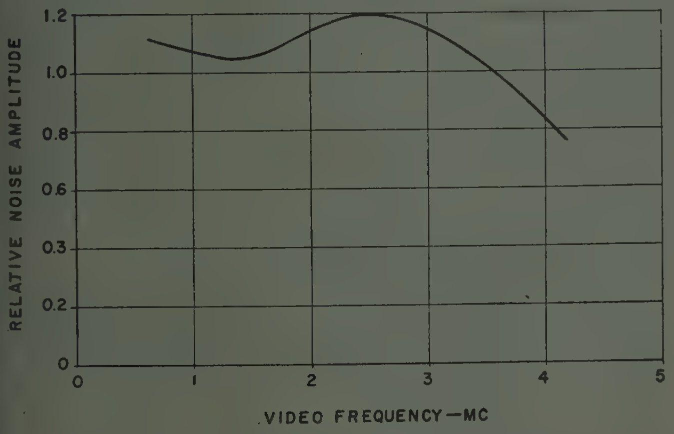

Fig. 3—Relative spectrum of noise generator.

The signal used during the tests was a standard NTSC color transmission, as specified in the NTSC press release of November 26, 1951, of two of the NTSC reference test slides which showed a girl's face next to a sunflower set against a blue-sky background and a vase of roses set against a black-and-white background containing considerable detail. The color sync contained nine cycles of color carrier whose peak-to-peak height was 90 per cent of the deflection-sync height.

A trinescope adjusted for a highlight brightness of 35 foot-lamberts was used for the display of the composite picture. To avoid bias by the observer in rating the picture, the functions of the various controls and their calibration was disguised so that observer comment was restricted to the picture itself.

Studies by Mertz, Fowler, and Christopher[4] have shown that certain standard comments have an approximately linear relationship to psychological or liminal units. Accordingly, their recommended comments were used in the tests as rating factors, i.e., the observers were asked to adjust the controls until a particular comment was obtained. These impairment rating comments were:

1. Not perceptible.
2. Just perceptible.
3. Definitely perceptible, but only slight impairment to picture.
4. Impairment to picture but not objectionable.
5. Somewhat objectionable.
6. Definitely objectionable.
7. Not usable.

Ten observers at a viewing distance of four times the picture height used these comments to rate the thermal-noise performance of the picture, brightness information, the picture-chroma information, deflection synchronization, frame synchronization, and color synchronization of the color receiver on two different pictures. In addition, noise was injected into all channels of a color and of a monochrome receiver to show total picture impairment under the same noise conditions.

The averages of the noise settings at each impairment rating are plotted in Fig. 4 for individual functions and in Fig. 5 for total performance and for color synchronization. Fig. 4 shows that the limiting factor in the noise performance of the color signal is the luminosity information at all noise levels. Second-order limitation is primarily the deflection synchronization though the chroma-picture information has about the same noise performance. Thus, it may be seen that the controlling factors in the color-system performance are exactly the same as those met in the basic monochrome signal.

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[4] Pierre Mertz, A. D. Fowler, and H. N. Christopher, "Quality rating of television images," Proc. I.R.E.; November, 1950.

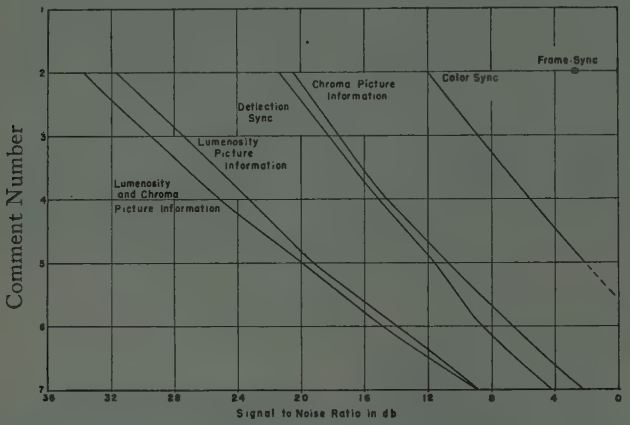

Fig. 4—Subjective response to noise in color receiver functions.

The chroma-picture information showed about the same impairment as deflection synchronization at low noise levels, but deflection synchronization caused appreciably more impairment under high noise conditions. It is interesting to note the noise-performance differential that exists between the picture luminosity and chroma information which demonstrates the noise advantages obtained by the constant-luminance features of the NTSC system.

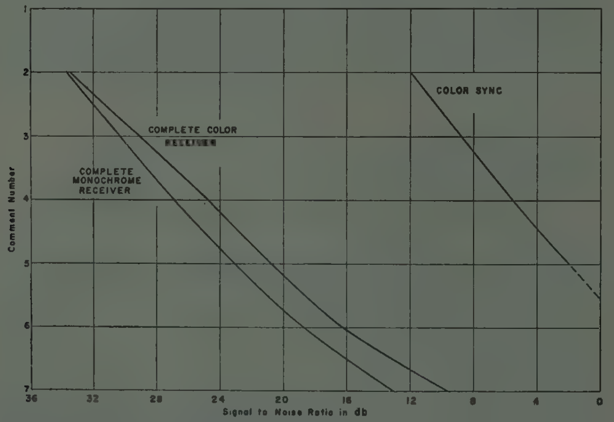

Fig. 5—Subjective response to noise in color and monochrome receivers.

The impairment caused by color synchronization was far lower than that caused by the picture information and deflection synchronization. In fact, the picture information had reached a point where it was classified as definitely objectionable before the color-sync impairment became just perceptible. When the picture information has reached a not usable classification the color-sync degradation is still so low that the picture would be entirely satisfactory even with a noise-free picture. The impairment that is obtained could not be distinguished under normal circumstances where noise is applied to all receiver functions simultaneously. The technique used in testing in effect acts as a magnifier upon impairment.

Several reasons exist for the excellent performance of the color-sync circuits. Most important is the high-energy content of the burst, the ease with which it may be separated from general-noise disturbances at frequencies, times, or amplitudes other than those corresponding to the burst period, frequency and amplitude range, and the fact that little information is required to operate the circuits permitting the use of stiff memory-protective circuits. Second, the NTSC color system inherently minimizes disturbances in the color functions including color synchronization by use of constant-luminance transmission. This means that any fluctuation of the color-reference signal that may exist will have no effect upon luminance, but only hue or saturation to which the eye is relatively insensitive for short-time disturbances as is shown in the noise-performance differential between luminosity and chromaticity in Fig. 4.

Furthermore, the color-sync circuit used was by no means representative of ultimate performance. It is believed to be representative of ordinary low-cost AFC circuits using a conventional LC oscillator and utilizing a total of three tubes which includes the burst-separator tube. The unit had a lock-in range of plus or minus 5,000 cycles about the color-carrier frequency, which was maintained with ten-to-one changes of circuit-input levels. Lock-in time was less than one-tenth second from a frequency of 2,000 cycles off the reference. The maximum-phase shift from the nominal-center value was 6° which corresponded to a free-running oscillator frequency 500 cycles off the nominal value.

The nature of the disturbance in the picture caused by high noise in the color synchronization is a slight fluctuation or flutter of the instantaneous hue and saturation. Since the noise is random, the disturbance tends to average out so that little apparent change is observable. In particular, the disturbance has little subjective weight in comparison with a function such as deflection synchronization.

In Fig. 5 it may be noted that the subjective response to an over-all color receiver was more tolerant under thermal-noise interference than to the monochrome receiver. In general, this should be expected since the color receiver provides far more useful information; yet, as Fig. 4 shows, the impairment exists primarily in only the monochrome functions, leaving the added information relatively unimpaired. Again, this performance differential shows that the color functions, including color sync, are relatively unaffected by noise.

To find the influence of amplitude variations upon color-synchronizing performance, subjective tests were run with varied burst amplitude in the transmitted signal as a function of the video signal-to-noise ratio. This condition could be encountered when the effective antenna selectivity is great enough to reduce the higher-video frequencies. As shown in Fig. 6, the operation is still adequate down to quite low-burst amplitudes. In general, the same type of curve is obtained when fewer burst cycles are transmitted—not becoming objectionable until the burst contains less than 2 cycles and the signal-to-noise ratio lower than 2 to 3 db.

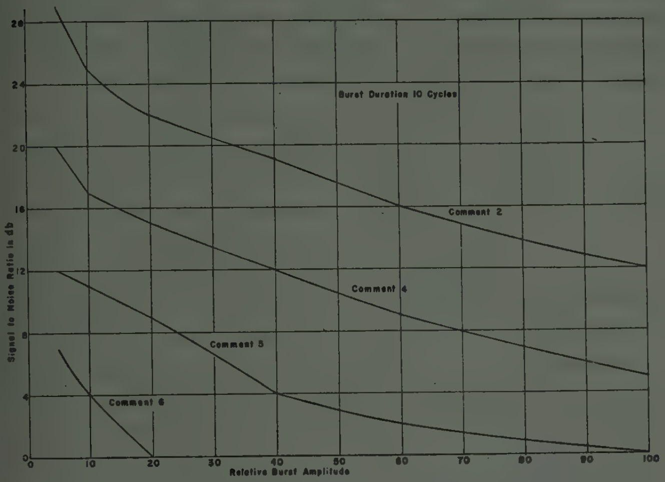

Fig. 6—Influence of burst amplitude on color synchronizing performance.

It is also found that high noise conditions in the picture information cause the eye to be even more tolerant of changes in the color-reference phase. This effect is shown in Fig. 7, which is a plot of the range of allowable phase variation of the color-reference signal for a given signal-to-noise ratio. Apparently, the noise disturbance in the picture obscures the hue disturbance sufficiently so that the eye is relatively unaware of the error.

The results shown here are probably somewhat pessimistic since the tests used slides for program material. In practice, changing program content and dramatic interest may decrease the observer's ability to detect given amounts of noise impairment. Although the subjective judgment of only ten observers does not give a universal result, it is believed that these studies show that the NTSC color-synchronization system is adequate from the standpoint of thermal-noise performance. Extensive field experience in NTSC and Philco tests with weak-signal performance have further strengthened these conclusions. Field experience with severe impulse noise has shown no signs of deterioration other than deflection-synchronizing functions, and therefore, the effect is relatively unimportant in color-synchronizing considerations. Furthermore, theoretical considerations[5] agree closely with the experimental results, indicating that the NTSC color-synchronizing specifications are completely satisfactory and adequate. Indeed, the studies as shown in Figs. 4 and 5 indicate that the problem of noise in the NTSC color system is primarily the same as in monochrome-improving receiver noise figures and noise immunity of deflection-synchronization circuits.

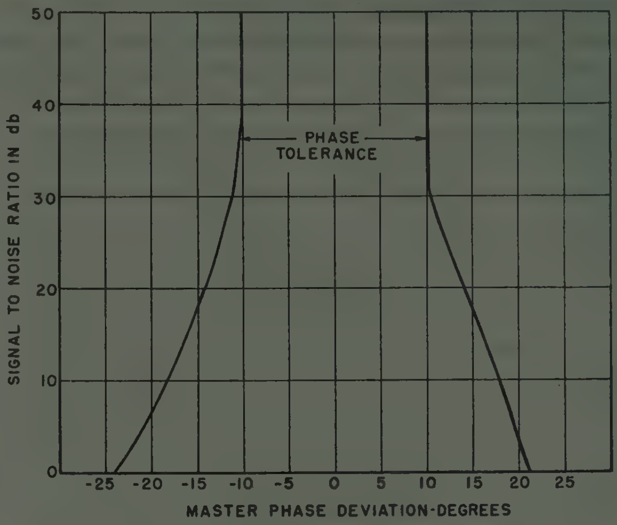

Fig. 7—Influence of picture noise on phase tolerances.

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[5] T. S. George, "Performance of carrier synchronizing systems," Proc. IRE; February, 1951.

Measurement of the Small Signal Parameters of Transistors*

GEOFFREY KNIGHT, JR.†, RICHARD A. JOHNSON†,
AND ROLAND B. HOLT†, SENIOR MEMBER, IRE

***Summary*—A theoretical study of various small signal parameters of transistors shows that the set of parameters which is most appropriate for the description of the circuit operation of junction transistors differs from the set which is customarily measured for point contact transistors. A uniform method for measuring both sets of parameters is described in detail and the sources of error inherent in this method are discussed.**

INTRODUCTION

IN THE DESIGN of vacuum tube amplifiers and related circuits it is customary to represent the tube by an equivalent circuit, the components of which include such differential properties of the tube as its transconductance, its amplification factor, and its plate resistance. This makes it possible to treat the tube as a network of linear circuit elements which greatly simplifies the analysis of circuits in the region of

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† Transistor Products, Inc., Brighton, Massachusetts.

the tube operation where the linear approximation is valid. Similarly, in the design of many kinds of transistor circuits it is appropriate and convenient to represent the transistor by an equivalent circuit composed of linear elements, the values of which are related to the differential properties of the transistor.

The term differential property refers to the rate at which a dependent variable changes as one of the independent variables is changed while the other independent variable is held constant. These differential properties are often called small-signal parameters. One of the most significant of the transistor parameters is the short-circuit current gain, denoted by α, which corresponds in some ways to the open-circuit voltage gain, μ, of a vacuum tube. Another parameter of interest, r_{22}, is very similar to the plate resistance of a tube. On the other hand, the transconductance of a transistor is not a particularly interesting parameter whereas its transresistance is important. This difference stems from the fact that a transistor is essentially a current operated device while a vacuum tube is a voltage operated device. For a vacuum tube triode with a negative grid-to-cathode voltage there are only three variables, two independent and one dependent, since the grid current is always zero. As a result only two of the three triode parameters are independent, anyone of them being expressible as a product or ratio of the other two. For a transistor, however, there are always four variables, two independent and two dependent, and as a result four independent parameters must be determined in order to completely specify the response of a transistor as a linear circuit element. It is the purpose of this paper to define and discuss the parameters of interest for transistor circuit analysis and to describe a particular method of measuring them.

Discussion of the Small-Signal Parameters

1. r_{11}, r_{12}, r_{21}, r_{22}.

For purposes of circuit analysis a transistor can be considered to be an active four terminal network.[1] For any such network the voltages across any two pairs of terminals and the currents flowing at those pairs of terminals are functionally related and these functions completely characterize the performance of the network. One of six possible ways to express this functional relationship is shown in (1) and (2) where capital letters designate constant values of voltage and current.

$$V_1 = V_1(I_1, I_2) \tag{1}$$

$$V_2 = V_2(I_1, I_2). \tag{2}$$

The curves obtained by plotting these functions with one current as the independent variable and the other as a parameter are known as characteristic curves of the network.

If the currents I_1 and I_2 make low frequency excursions about some bias values I_{10} and I_{20} as they do when a low frequency alternating signal is applied to the network, the functions V_1 and V_2 both have the form

$$V(t) = V_0 + v(t) = V[I_{10} + i_1(t), I_{20} + i_2(t)] \tag{3}$$

where $V_0 \equiv V(I_{10}, I_{20})$ and $v(t)$ and $i(t)$ are the instantaneous values of the voltage and current excursions. This function can be expanded in a Taylor's series as

$$\begin{aligned} V(t) &= V_0 + v(t) \\ &= V(I_{10}, I_{20}) + \frac{\partial V}{\partial I_1}\cdot i_1(t) + \frac{\partial V}{\partial I_2}\cdot i_2(t) \\ &\quad + \frac{1}{2}\left[\frac{\partial^2 V}{\partial I_1^2}\cdot i_1^2(t) + \frac{\partial^2 V}{\partial I_2^2}\cdot i_2^2(t) \right. \\ &\quad \left. + 2\,\frac{\partial^2 V}{\partial I_1 \partial I_2}\cdot i_1(t) i_2(t)\right] + \cdots \end{aligned} \tag{4}$$

where the partial derivatives are evaluated at $I_1 = I_{10}$ and $I_2 = I_{20}$.

If the current excursions $i_1(t)$ and $i_2(t)$ are so small that all the terms of this expansion of higher order than the first can be neglected, it reduces to the form:

$$v_1(t) = r_{11}i_1(t) + r_{12}i_2(t) \tag{5}$$

$$v_2(t) = r_{21}i_1(t) + r_{22}i_2(t) \tag{6}$$

where

$$\begin{aligned} r_{11} &\equiv \left[\frac{\partial V_1}{\partial I_1}\right]_{I_2=\text{const.}} & r_{12} &\equiv \left[\frac{\partial V_1}{\partial I_2}\right]_{I_1=\text{const.}} \\ r_{21} &\equiv \left[\frac{\partial V_2}{\partial I_1}\right]_{I_2=\text{const.}} & r_{22} &\equiv \left[\frac{\partial V_2}{\partial I_2}\right]_{I_1=\text{const.}} \end{aligned} \tag{7}$$

are the slopes of the set of characteristic curves implicit in (1) and (2).

The linear equations (5) and (6) correspond to the circuit shown in Fig. 1. It is apparent from this figure that r_{11} is the open circuit (*i.e.* $i_2 = 0$) input resistance and that r_{22} is the open circuit (*i.e.* $i_1 = 0$) output resistance of the network. It is also apparent that r_{12} is a feedback resistance whereby a current i_2 flowing in the output circuit generates a voltage $r_{12}i_2$ in the input circuit, and the r_{21} is a forward transfer resistance whereby a current i_1 flowing in the input circuit generates a voltage $r_{21}i_1$ in the output circuit.

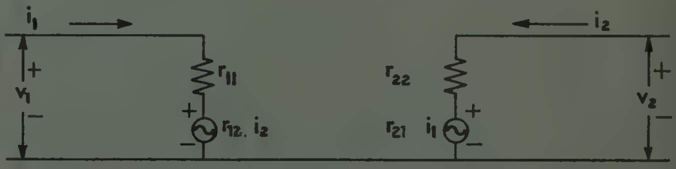

Fig. 1—Equivalent circuit of the network described by equations (5) and (6).

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[1] R. M. Ryder and R. J. Kircher, "Some circuit aspects of the transistor," *Bell System Tech. Jour.*, vol. 28, pp. 367–401; July, 1949.

2. g_{11}, γ, α, r_{22}.

In the previous discussion we have considered the input and output currents to be the natural independent variables of the network. However, in some ways it is more appropriate to the nature of a transistor to consider the input current and the output voltage to be the independent variables in the usual circuit connections. This is true from the point of view of the physics of the transistor because the parameter which is the most direct measure of transistor action, namely the current gain, is defined in terms of an input current signal and a constant output voltage. It is also true from the point of view of measurement and of circuit operation because the transistor is inherently a low input resistance, high ouput resistance device in the most useful circuit connections. In fact, the input resistance of a transistor may be so low that the input voltage cannot be conveniently controlled whereas the output resistance may be so high that the output current cannot be conveniently controlled. As an example we may consider the measurement of r_{11}, r_{12}, r_{21}, and r_{22} by the method to be discussed later in this paper. In each case the dynamic resistance of the collector dc supply must be very much greater than r_{22} if the measurement is to be accurate. However, the value of r_{22} for a junction transistor may be many megohms so the resistance of the supply would have to be of the order of hundreds of megohms, which is impossible to achieve at any reasonable measuring frequency because of the effects of distributed capacity.

When I_1 and V_2 are taken to be the independent variables, the functional relationship between the network voltages and currents may be written:

$$V_1 = V_1(I_1, V_2) \tag{8}$$

$$I_2 = I_2(I_1, V_2). \tag{9}$$

If the current I_1 and the voltage V_2 make sufficiently small excursions about some bias values, the Taylor's series expansions of these equations reduce to the linear relations:

$$v_1 = \frac{1}{g_{11}} i_1 + \gamma v_2 \tag{10}$$

$$i_2 = -\alpha i_1 + \frac{1}{r_{22}} v_2 \tag{11}$$

where

$$\frac{1}{g_{11}} \equiv \left[\frac{\partial V_1}{\partial I_1}\right]_{V_2=\text{const.}}, \quad \gamma \equiv \left[\frac{\partial V_1}{\partial V_2}\right]_{I_1=\text{const.}},$$
$$\alpha \equiv -\left[\frac{\partial I_2}{\partial I_1}\right]_{V_2=\text{const.}} \quad \frac{1}{r_{22}} \equiv \left[\frac{\partial I_2}{\partial V_2}\right]_{I_1=\text{const.}} \tag{12}$$

Thus g_{11}, γ, α, and r_{22} are the natural parameters when I_1 and V_2 are the independent variables. This set of parameters has recently been adopted by Bell Telephone Laboratories and Western Electric Company for the specifications of their junction transistors. (Private communication.) Also, these parameters are measured for junction transistors by a test set manufactured by Transistor Products, Inc.

The linear equations (10) and (11) correspond to the circuit shown in Fig. 2. Examination of this circuit shows that g_{11} is the short circuit ($v_2=0$) input conductance and that r_{22} is the open circuit ($i_1=0$) output resistance. It is also apparent that γ is a voltage feedback ratio such that if a voltage v_2 appears across the output terminals, a voltage generator of magnitude γv_2 appears in the input circuit. The final parameter, α, is seen to be the short circuit ($v_2=0$) current gain of the network such that if a current i_1 flows in the input circuit a current generator $-\alpha i_1$ appears in the output circuit. The parameters g_{11}, γ, and α are related to r_{11}, r_{12}, r_{21}, and r_{22} by the expressions

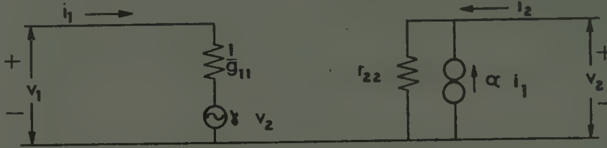

Fig. 2—Equivalent circuit of the network described by (10) and (11).

$$1/g_{11} = r_{11} - \alpha r_{12}, \quad \gamma = r_{12}/r_{22}, \quad \text{and} \quad \alpha = r_{21}/r_{22}.$$

The values of r_{11}, r_{12}, r_{21}, and r_{22} and of g_{11}, α, and γ for a given transistor depend upon which lead of the transistor is taken to be the common terminal (usually called the ground terminal) for the voltages v_1 and v_2. In this paper we shall only find it necessary to consider parameters which are associated with the grounded base connection.

In general the characteristic curves of a transistor are nonlinear so the values of the parameters must be determined as functions of the operating point, e.g. $r_{11}(I_{10}, I_{20})$ and $\alpha(I_{10}, V_{20})$, if the small signal response of the transistor at low frequencies is to be completely specified.

As the signal frequency increases, capacitative and transit time effects in the transistor become more important so that it becomes necessary to consider impedances and admittances rather than resistances and conductances. However, in all the measurements described here, signal frequency will be assumed to be so low that frequency dependent effects can be ignored.

MEASUREMENT OF THE SMALL SIGNAL PARAMETERS

A standard method of measuring the parameters r_{11}, r_{12}, r_{21}, and r_{22} is based on (5) and (6).[1,2] They may be

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[2] K. Lehovec, "Testing transistors," *Electronics*, vol. 22, pp. 88–89; June, 1949.

rewritten for the grounded base connection:

$$v_{eb} = r_{11b}i_e + r_{12b}i_c \tag{13}$$

$$v_{cb} = r_{21b}i_e + r_{22b}i_c \tag{14}$$

where v_{eb} and v_{cb} are the signal voltages between emitter and base and collector and base respectively and i_e and i_c are the signal currents into emitter and collector.

As an example, let us consider the case of r_{11b}. Solution of (13) for r_{11b} gives

$$r_{11b} = \frac{v_{eb}}{i_e} - r_{12b}\frac{i_c}{i_e}. \tag{15}$$

Thus r_{11b} equals the ratio v_{eb}/i_e if the second term is so small compared with the first that it can be neglected. This will be the case if

$$i_c \ll -\frac{r_{11b}}{r_{12b}}i_e,$$

that is, if the magnitude of the collector current is very small compared with the emitter current. Now if a load resistance r_L is connected between collector and base so that $V_{cb} = -r_L i_c$, we have (from (14))

$$i_c = \frac{-r_{21b}}{r_{22b} + r_L}i_e$$

so if

$$\frac{r_{21b}}{r_{22b} + r_L} \ll \frac{r_{11b}}{r_{12b}},$$

i.e. if

$$r_L \gg \frac{r_{12b}}{r_{11b}}r_{21b},$$

the required condition will be established.

Similar relations can be derived for r_{12b}, r_{21b}, and r_{22b}. The method, then, consists of introducing a small alternating signal current i_e or i_c; establishing a condition, $i_c = 0$ or $i_e = 0$ by means of an external resistance; and measuring a small signal voltage. The value of the parameter is given by the ratio of the output voltage to the input current. Therefore, if the input current is made equal to unity, the value of the parameter is numerically equal to the output voltage.

This method can also be used to measure $1/g_{11b}$, γ_b, α_b, $1-\alpha_b$,[3] and $1/r_{22b}$. The basic equations, obtained by rewriting (10) and (11) for the grounded base connection and by using the relation $i_b = -(i_e + i_c)$, are:

$$v_{eb} = \frac{1}{g_{11b}}i_e + \gamma_b v_{cb} \tag{16}$$

$$i_c = -\alpha_b i_e + \frac{1}{r_{22b}}v_{cb} \tag{17}$$

$$i_b = -(1-\alpha_b)i_e - \frac{1}{r_{22b}}v_{cb} \tag{18}$$

These equations can be solved so as to express each of the parameters in terms of a ratio of output signal to input signal plus a term which can be made negligibly small by a suitable resistance in the external circuit.

A summary of the type of signal required, the type of measurement to be made, the condition to be established, and the necessary values of external resistance for each of the desired parameters is given in Table I.

Because the required signals, conditions, and measurements are similar for all of the parameters, the appropriate measuring circuits all have similar components. In each case there must be a source of signal current or voltage, a source of dc current for the emitter, a source of dc current or voltage for the collector, and an ac millivoltmeter or milliammeter. The arrangements of these components which are required for the measurement of the parameters discussed in this paper are shown schematically in Figs. 3, 4, and 5.

In these circuits the symbols r_s, r_k, r_L, r_m, and r_n represent the dynamic resistances of the signal source, of the emitter dc supply, of the collector dc supply, of the millivoltmeter, and milliammeter respectively.

Sources of Error

There are inherent in this method of evaluating small signal parameters a number of sources of error which affect the accuracy of measurement.

First we may consider the error due to the nonlinearity of the characteristic curves which results when a finite rather than an infinitesimal signal is used to measure the slope. Let us take the case of r_{11} as an example. If the input signal current is a pure sine wave, $i_1(t) = i_{10}\sin\omega t$, while the other current, $i_2(t)$, equals zero; (4) becomes:

$$v_1(t) = \frac{\partial V_1}{\partial I_1}i_{10}\sin\omega t + \frac{1}{2}\frac{\partial^2 V_1}{\partial I_1^2}i_{10}^2\sin^2\omega t + \frac{1}{6}\frac{\partial^3 V_1}{\partial I_1^3}i_{10}^3\sin^3\omega t + \cdots \tag{19}$$

where the second, third, and higher terms vanish compared with the first only if i_{10} is infinitesimal or if V_1 is a linear function of I_1.

In order to determine the order of magnitude of the error introduced by these additional terms it is necessary to consider the nature of the measurement of $v_1(t)$. If the signal voltage to be measured passes through an amplifier which is sharply tuned to the measuring frequency before it reaches the voltmeter, the second harmonic of the signal frequency will be greatly attenuated and higher order harmonics will be effectively eliminated

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[3] In the case of junction transistors the value of α_b may be so close to unity that it is desirable to measure $1-\alpha_b$.

TABLE I

SUMMARY OF THE REQUIREMENTS FOR THE MEASUREMENT OF TRANSISTOR PARAMETERS BY THE METHOD DESCRIBED IN THE TEXT

				Required Values of Internal Resistance			
Parameter	Signal	Measurement	Condition	Signal Source	Emitter DC Supply	Collector DC Supply	Circuit Diagram
$r_{11b}=\frac{v_{eb}}{i_e}-r_{12b}\frac{i_c}{i_e}$	i_e	v_{eb}	$i_c\ll\frac{r_{11b}}{r_{12b}}i_e$	$r_s\gg r_{11b}$	$r_k\gg r_{11b}$	$r_L\gg\frac{r_{12b}}{r_{11b}}r_{21b}$	3
$\frac{1}{g_{11b}}=\frac{v_{eb}}{i_e}-\gamma_b\frac{v_{cb}}{i_e}$	i_e	v_{eb}	$v_{cb}\ll\frac{i_e}{\gamma_b g_{11b}}$	$r_s\gg\frac{1}{g_{11b}}$	$r_k\gg\frac{1}{g_{11b}}$	$r_L\ll\frac{1}{\alpha_b\gamma_b g_{11b}}$	3
$r_{12b}=\frac{v_{eb}}{i_c}-r_{11b}\frac{i_e}{i_c}$	i_c	v_{eb}	$i_e\ll\frac{r_{12b}}{r_{11b}}i_c$	$r_s\gg r_{22b}$ $r_s\gg r_{11b}$	$r_k\gg r_{11b}$ $r_k\gg r_{11b}$	$r_L\gg r_{22b}$ $r_L\gg r_{22b}$	3 3
$r_{21b}=\frac{v_{cb}}{i_e}-r_{22b}\frac{i_c}{i_e}$	i_e	v_{cb}	$i_c\ll\alpha_b i_e$	$r_s\gg r_{22b}$	$r_k\gg\alpha_b r_{12b}$	$r_L\gg r_{22b}$	3
$r_{22b}=\frac{v_{cb}}{i_c}-r_{21b}\frac{i_e}{i_c}$	i_c	v_{cb}	$i_e\ll\frac{1}{\alpha_b}i_c$	$r_s\gg\frac{1}{g_{11b}}$	$r_k\gg\frac{1}{g_{11b}}$	$r_L+r_n\ll r_{22b}$	4
$\alpha_b=-\frac{i_c}{i_e}+\frac{1}{r_{22b}}\frac{v_{cb}}{i_e}$	i_e	i_c	$v_{cb}\ll r_{21b}i_e$	$r_s\gg\frac{1}{g_{11b}}$	$r_k\gg\frac{1}{g_{11b}}$	$r_L-\frac{1-\alpha_b}{\alpha_b}r_n$	4
$1-\alpha_b=-\frac{i_b}{i_e}-\frac{1}{r_{22b}}\frac{v_{cb}}{i_e}$	i_e	i_b	$v_{cb}\ll(1-\alpha_b)r_{22b}i_e$			$\ll\frac{1-\alpha_b}{\alpha_b}r_{22b}$	
$\gamma_b=\frac{v_{eb}}{v_{cb}}-\frac{1}{g_{11b}}\frac{i_e}{v_{cb}}$	v_{cb}	v_{eb}	$i_e\ll g_{11b}\gamma_b v_{cb}$	$r_s\ll r_{22b}$	$r_k\gg\frac{1}{g_{11b}}$		5
$\frac{1}{r_{22b}}=\frac{i_c}{v_{cb}}+\alpha_b\frac{i_e}{v_{cb}}$	v_{cb}	i_c	$i_e\ll\frac{v_{cb}}{\alpha_b r_{22b}}$	$r_s+r_n\ll r_{22b}$	$r_k\ll\alpha_b\gamma_b r_{22b}$		5

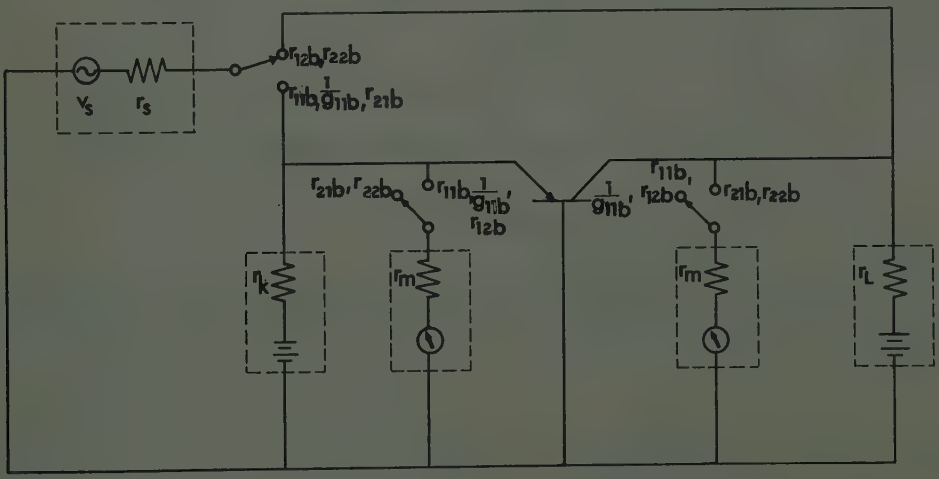

Fig. 3—Prototype circuit for the measurement of the parameters r_{11b}, $1/g_{11b}$, r_{12b}, r_{21b}, r_{22b}.

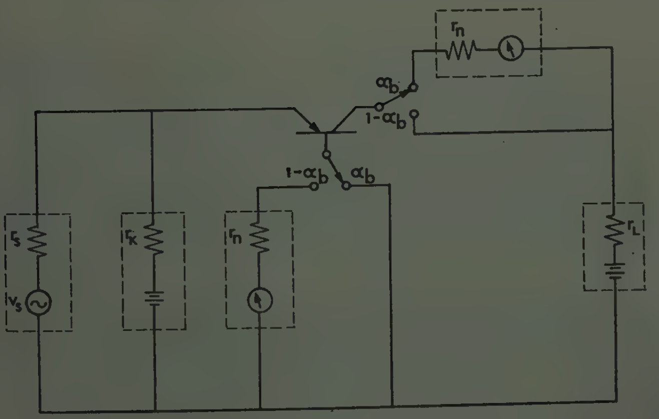

Fig. 4—Prototype circuit for measurement of parameters α_b, $1-\alpha_b$.

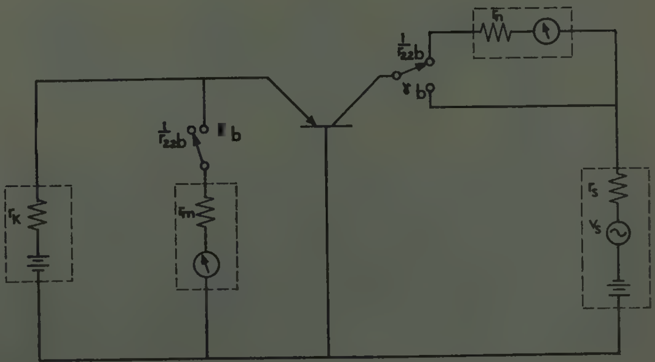

Fig. 5—Prototype circuit for measurement of parameters $1/r_{22b}$, γ_b.

Thus if the input to the amplifier is

$$v_1(t) = \left(\frac{\partial V_1}{\partial I_1} i_{10} + \frac{1}{8}\frac{\partial^3 V_1}{\partial I_1^3} i_{10}^3 + \cdots\right)\sin \omega t$$

$$+ \frac{1}{4}\frac{\partial^2 V_1}{\partial I_1^2} i_{10}^2 \cdot (1 - \cos 2\omega t)$$

$$- \frac{1}{24}\frac{\partial^3 V_1}{\partial I_1^3} i_{10}^3 \sin 3\omega t + \cdots \qquad (20)$$

(which can be obtained from (19) by trigonometric identity), the output from the amplifier is proportional to

$$\left(\frac{\partial V_1}{\partial I_1} i_{10} + \frac{1}{8}\frac{\partial^3 V_1}{\partial I_1^3} i_{10}^3 + \cdots\right)\sin \omega t \qquad (21)$$

plus a small second harmonic term which is further reduced by the averaging action of the meter. If the signal is infinitesimal, i_{10} being so small that the second term of this expression is completely negligible compared with the first, the correct value, $r_{11} = \partial V_1/\partial I_1$, is obtained. When a finite signal is used, the second term may not be negligible if the rate of change of curvature of the characteristic, $\partial^3 V_1/\partial I_1^3 = \partial^2 r_{11}/\partial I_1^2$, is appreciable. The fractional error in the measurement of r_{11} due to the combined effects of finite signal size and curvature of the characteristic is to an excellent approximation the ratio of the second (error) term of (21) to the first (correct) term, i.e.

$$\frac{1}{8}\frac{1}{r_{11}}\frac{\partial^2 r_{11}}{\partial I_1^2} i_{10}^2 \qquad (22)$$

where i_{10} is the amplitude of the input signal current. An expression similar to (21) can be derived for each of the other parameters so in each case the fractional error is similar in form to (22). The general expression for the fractional error in the measured value, p, of a parameter is either

$$\frac{1}{8}\frac{1}{p}\frac{\partial^2 p}{\partial I_s^2} i_{s0}^2 \quad \text{or} \quad \frac{1}{8}\frac{1}{p}\frac{\partial^2 p}{\partial V_s^2} v_{s0}^2, \qquad (23)$$

depending on whether the signal is a current i_s or a voltage v_s.

It is apparent that this error can be made negligibly small by making i_{s0} or v_{s0} sufficiently small. However, the presence of transistor and thermal noise establishes a lower limit to the useful size of the signal. Since transistors are relatively noisy, this may be an important consideration.

A second source of error associated with the signal current results from the finite internal resistance of the signal source. In measuring any of the parameters the magnitude of the input signal is assumed to be independent of the resistance of the transistor, but this will only be true if the source resistance is either very much greater or very much less than the input resistance of the transistor. The order of magnitude of the fractional error in the measured value of a parameter is either the ratio of the input resistance of the transistor to the internal resistance of the source or its reciprocal, depending on whether a current source or a voltage source is required.

A third source of error associated with the signal current results from the shunting of the signal through the voltmeter and through the dc supply which is in parallel with the signal source since the part of the input signal which bypasses the transistor in this way cannot contribute to the output signal. In practice the resistance of the voltmeter can be made so large that its shunting effect can be neglected. The order of magnitude of the fractional error due to the shunting of the dc supply is the ratio of the input resistance of the transistor to the dynamic resistance of the dc supply.

The error associated with establishing the necessary condition can be illustrated by the case of r_{11b}. The magnitude of the resistance r_L required to make the second term of (15) negligible can be determined in the following way: The equations for the flow of collector current in the circuit are:

$$v_{cb} = r_{21b} i_e + r_{22b} i_c = - r_L i_c \qquad (24)$$

so that

$$\frac{i_c}{i_e} = - \frac{r_{21b}}{r_L + r_{22b}} \cong - \frac{r_{21b}}{r_L} \qquad (25)$$

if $r_L \ll r_{22b}$. Substitution of this result into (15) gives:

$$r_{11b} = \frac{v_{eb}}{i_e} + r_{12b}\frac{r_{21b}}{r_L}. \qquad (26)$$

The fractional error which results if the second term is neglected in the computation of r_{11b} is

$$\frac{r_{12b}}{r_{11b}} \cdot \frac{r_{21b}}{r_L}.$$

Therefore, for this error to be small, r_L must be many times greater than

$$\frac{r_{12b}}{r_{11b}} r_{21b}.$$

For example, for a point contact transistor with $r_{21b} = \alpha_b r_c = 50K$ and $r_{12}/r_{11} = \frac{1}{2}$ this error will be $<1\%$ only if $r_L > 2.5$ megohms. Similar expressions giving the error involved in the assumption that a parameter equals the ratio of the output signal to the input signal can be derived in each case.

The values of r_s, r_k, and r_L which are required if these errors are to be kept small are given in Table I for each parameter. The approximate values of the fractional error in the measurement of a given parameter due to too small or too large a value of these internal resistances is given by the ratio of the value of the parameter in question to the number in the 5th, 6th, and 7th columns. For example, the approximate value of the fractional error which would result in the measurement of r_{11b} if the internal resistance of the source were finite would be r_{11b}/r_s.

It should also be noted that there will necessarily be some stray capacity between emitter and base and between collector and base which will be in shunt with the dc supplies. At a given signal frequency the minimum realizable values of these capacities will set an upper limit on the useful values of r_k and r_L. If $C \sim 10\mu\mu$f, this limit is ~ 150 megohms for a signal frequency of 100 cps; ~ 3 megohms for a signal frequency of 5 kc.

The final errors to be considered are those involved in setting the input current or voltage and in measuring the output voltage or current. By using the same metering circuit to perform both these operations so that the measurement of the output signal is reduced to a comparison of the unknown resistance with a standard resistance or the unknown ratio with a standard resistance ratio it is possible to eliminate any errors due to the absolute calibration of the meter.

APPENDIX

Formulas for such circuit properties of the transistor as its input and output resistances and its power gain are generally given in terms of the equivalent circuit resistances r_e, r_b, r_c and r_m[4-5] which can be readily computed from the measurable parameters (cf. Table II). Reference 5 contains an extensive list of such formulas.

TABLE II

EQUIVALENT-CIRCUIT PARAMETERS OF TRANSISTORS IN TERMS OF THE MEASURED PARAMETERS

$$\begin{array}{ll}
r_e = r_{11b} - r_{12b} = 1/g_{11b} - (1-\alpha_b)\gamma_b r_{22b} & \\
r_b = r_{12b} \qquad = \gamma_b r_{22b} & \\
r_c = r_{22b} - r_{12b} = (1-\gamma_b) r_{22b} & \cong r_{22b} \\
r_m = r_{21b} - r_{12b} = (\alpha_b - \gamma_b) r_{22b} & \cong \alpha_b r_{22b} \\
a = \dfrac{r_{21b} - r_{12b}}{r_{22b} - r_{12b}} = \dfrac{\alpha_b - \gamma_b}{1-\gamma_b} & \cong \alpha_b
\end{array}$$

The approximations are excellent because $\gamma_b \ll 1$, α_b for a satisfactory transistor.

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[4] R. L. Wallace, Jr. and W. J. Pietenpol, "Some circuit properties and applications of n-p-n transistors," PROC. I.R.E., vol. 39, pp. 753–767; July, 1951.

[5] F. R. Stansel, "Transistor Equations," *Electronics*, vol. 26, No. 3, pp. 156, 158; March, 1953.

A Note on the Analysis of Vacuum Tube and Transistor Circuits*

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***Summary*—A simple method is described for setting up the node (or mesh) equations for vacuum tube and transistor circuits. The method is based on the use of admittance (or impedance) coefficients tables for the vacuum tube and the transistor. It does not involve the replacement of the vacuum tube and the transistor by their equivalent circuits, and hence does not require a redrawing of the circuit diagram. With the aid of the tables, the node (or mesh) equations are set up directly from inspection of the circuit diagram and without the use of any intermediate steps.**

I. INTRODUCTION

THE ANALYSIS of linear vacuum tube and transistor circuits is usually carried out in three, not necessarily separate, steps. First, each vacuum tube and transistor is replaced by an equivalent circuit consisting of one or more impedances and a voltage or current source; second, the node or mesh equations for the resulting network are set up as for a passive network, treating the voltage (or current) sources in the equivalent circuits as if they were known quantities; and third, the equivalent voltage (or current) source terms in the equations are replaced by expressions involving impressed voltages or node voltages or mesh currents or their combinations. These steps result in a system of linear equations which in the case of the node method of analysis are of the form

$$\begin{aligned}
J_1 &= Y_{11}V_1 + Y_{12}V_2 + \cdots + Y_{1n}V_n \\
J_2 &= Y_{21}V_1 + Y_{22}V_2 + \cdots + Y_{2n}V_n \\
&\cdots\cdots\cdots\cdots\cdots\cdots \\
J_n &= Y_{n1}V_1 + Y_{n2}V_2 + \cdots + Y_{nn}V_n,
\end{aligned} \tag{1}$$

where the J's are impressed node currents, the V's are node voltages and the Y's are admittance coefficients.

The object of this note is to outline a simpler method of setting up the node (or mesh) equations for linear vacuum tube and transistor circuits. The method makes use of tables of admittance (or impedance) coefficients for the vacuum tube and the transistor, and its application requires a minimum of thought and labor.

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Although the basic idea underlying the method is well known,[1–3] it appears that a description incorporating tables of admittance and impedance coefficients for the vacuum tube and the transistor (regarded as three-terminal elements) has not been given in the literature. Since experience with the method indicates it is considerably more effective than the conventional method, a brief exposition might be worthwhile.

II. Description of Method

For simplicity, the discussion will be limited to triodes and three-terminal transistors. Furthermore, only the node method of analysis will be considered in detail.

The method described here is illustrated in Fig. 1, which shows (a) a network N containing a three-terminal network Δ; (b) the network $N-\Delta$, which results from removing Δ from N; and (c) the network Δ. It is assumed that Δ is not inductively coupled to $N-\Delta$.

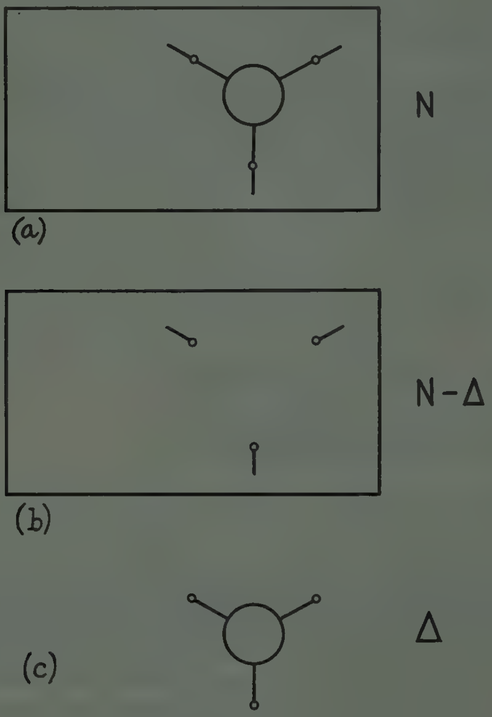

Fig. 1—(a) Active network N containing an active three-terminal element Δ. (b) Network N with Δ removed. (c) Active element Δ.

Let $Y_{\nu\mu(N)}$, $Y_{\nu\mu(N-\Delta)}$ and $Y_{\nu\mu(\Delta)}$ ($\nu, \mu = 1, 2, 3, \cdots, n$) be the admittance coefficients for N, $N-\Delta$ and Δ, respectively. From the definitions[4] of these admittances it follows at once that

$$Y_{\nu\mu(N)} = Y_{\nu\mu(N-\Delta)} + Y_{\nu\mu(\Delta)} \qquad (2)$$

with the understanding that $Y_{\nu\mu(\Delta)}$ is zero for all ν and μ which are not associated with the terminals of Δ.

Equation (2) may be used to reduce the determination of the $Y_{\nu\mu(N)}$ to two simpler problems, namely, the calculation of the $Y_{\nu\mu(N-\Delta)}$ and the calculation of the $Y_{\nu\mu(\Delta)}$. In general there is little or no advantage in using this reduction. However, when Δ is a standard network such as a vacuum tube or a transistor, the $Y_{\nu\mu(\Delta)}$ may be obtained from a table of admittance coefficients for Δ. In this way, the setting up of node equations for a network N containing one or more active elements Δ_1 $\Delta_2, \cdots, \Delta_m$, is reduced essentially to determining the admittance coefficients for the passive network $N-\Delta_1-\Delta_2-\cdots-\Delta_m$ (resulting from removing $\Delta_1, \Delta_2, \cdots, \Delta_m$ from N) and adding to these the corresponding admittance coefficients for $\Delta_1, \Delta_2, \cdots, \Delta_m$, the latter being obtained from tables such as those given below.

TABLE I

Admittance Coefficients Table for the Vacuum Tube

$Y_{gg}=0$	$Y_{gp}=0$	$Y_{gk}=0$
$Y_{pg}=g_m$	$Y_{pp}=g_p$	$Y_{pk}=-g_p-g_m$
$Y_{kg}=-g_m$	$Y_{kp}=-g_p$	$Y_{kk}=g_p+g_m$

where g, p and k represent the grid, plate and cathode terminals, respectively; g_m is the transconductance and $g_p=1/r_p$ = plate conductance.

TABLE II

Admittance Coefficients Table for the Transistor

$Y_{\epsilon\epsilon} = \dfrac{r_b+r_c}{r_\epsilon(r_b+r_c)+(1-\alpha)r_br_c}$	$Y_{\epsilon c} = \dfrac{-r_b}{r_\epsilon(r_b+r_c)+(1-\alpha)r_br_c}$	$Y_{\epsilon b} = \dfrac{-r_c}{r_\epsilon(r_b+r_c)+(1-\alpha)r_br_c}$
$Y_{c\epsilon} = \dfrac{-r_b-\alpha r_c}{r_\epsilon(r_b+r_c)+(1-\alpha)r_br_c}$	$Y_{cc} = \dfrac{r_b+r_\epsilon}{r_\epsilon(r_b+r_c)+(1-\alpha)r_br_c}$	$Y_{cb} = \dfrac{\alpha r_c-r_\epsilon}{r_\epsilon(r_b+r_c)+(1-\alpha)r_br_c}$
$Y_{b} = \dfrac{-r_c(1-\alpha)}{r_\epsilon(r_b+r_c)+(1-\alpha)r_br_c}$	$Y_{bc} = \dfrac{-r_\epsilon}{r_\epsilon(r_b+r_c)+(1-\alpha)r_br_c}$	$Y_{bb} = \dfrac{r_\epsilon+(1-\alpha)r_c}{r_\epsilon(r_b+r_c)+(1-\alpha)r_br_c}$

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[1] G. Kron, "Tensor Analysis of Networks," John Wiley and Sons, Inc., New York, N. Y.; 1949. In particular, note pp. 389–390.

[2] S. R. Deards, "Matrix theory applied to thermionic valve circuits," *Electronic Eng.*, vol. 24, pp. 264–277; June, 1952. In this paper tables of admittance and impedance coefficients are given for several basic vacuum tube circuits (regarded as two terminal-pair elements).

[3] J. Shekel, "Matrix representation of transistor circuits," Proc. I.R.E., vol. 40, pp. 1493–1497; November, 1952. The method given in this paper is closely related to that described in the present paper. However, the paper does not contain tables of impedance and admittance coefficients.

[4] $Y_{\nu\mu}$ is numerically equal to the current flowing toward the νth node through a short circuit between the νth node and ground when (a) all current sources are open-circuited, (b) all nodes except the μth node are connected to ground, and (c) a potential of one volt is impressed upon the μth node. In the case of a passive network which does not contain coupled coils, $Y_{\nu\mu}$ ($\nu \neq \mu$) is equal to the negative of the admittance of the branch connecting the nodes ν and μ; and $Y_{\nu\nu}$ is equal to the sum of the admittances of the branches connected to the νth node.

where ϵ, c and b represent the emitter, collector and base terminals respectively; r_ϵ, r_c, and r_b are the incremental emitter, collector and base resistances; and α is the current amplification factor.

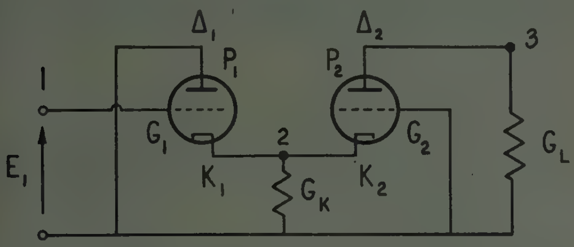

Fig. 2—Circuit analyzed in example 1.

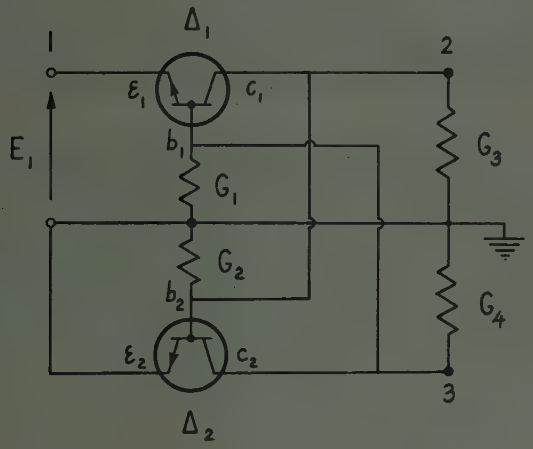

Fig. 3—Circuit analyzed in example 2.

As an illustration of the method, the node equations for the circuits shown in Figs. 2 and 3 will be set up. In the circuit shown in Fig. 2, the node voltage of node 1 is specified and hence the node equation for this node may be omitted. For nodes 2 and 3 the equations are of the form

$$0 = Y_{21}E_1 + Y_{22}V_2 + Y_{23}V_3 \tag{3}$$

$$0 = Y_{31}E_1 + Y_{32}V_2 + Y_{33}V_3. \tag{4}$$

The admittance coefficients in these equations are calculated in the following manner. Taking Y_{22}, for example, one has

$$Y_{22(N-\Delta_1-\Delta_2)} = G_k, \tag{5}$$

and from Table 1

$$Y_{22}(\Delta_1) = Y_{k_1k_1} = g_{p_1} + g_{m_1} \tag{6}$$

$$Y_{22}(\Delta_2) = Y_{k_2k_2} = g_{p_2} + g_{m_2}. \tag{7}$$

Thus

$$\begin{aligned} Y_{22(N)} &= Y_{22(N-\Delta_1-\Delta_2)} + Y_{22(\Delta_1)} + Y_{22(\Delta_2)} \\ &= G_k + g_{p_1} + g_{m_1} + g_{p_2} + g_{m_2}. \end{aligned} \tag{8}$$

Similar calculations for other coefficients yield the desired node equations

$$0 = -g_{m_1}E_1 + (G_k + g_{p_1} + g_{m_1} + g_{p_2} + g_{m_2})V_2 - g_{p_2}V_3 \tag{9}$$

$$0 = 0 - (g_{p_2} + g_{m_2})V_2 + (G_L + g_{p_2})V_3, \tag{10}$$

in which the lower case symbols represent quantities obtained from Table I.

Consider now the network shown in Fig. 3, which will be recognized as a simplified circuit of a negative-impedance converter. On inspecting the circuit diagram and making use of Table II, one readily obtains the following node equations

$$\begin{aligned} 0 = &- \frac{(r_b + \alpha r_c)E_1}{r_\epsilon(r_b + r_c) + (1 - \alpha)r_br_c} \\ &+ \left[G_2 + G_3 + \frac{2r_\epsilon + r_b + (1 - \alpha)r_c}{r_\epsilon(r_b + r_c) + (1 - \alpha)r_br_c}\right]V_2 \\ &+ \frac{(\alpha r_c - 2r_\epsilon)V_3}{r_\epsilon(r_b + r_c) + (1 - \alpha)r_br_c} \end{aligned} \tag{11}$$

$$\begin{aligned} 0 = &- \frac{r_c(1 - \alpha)E_1}{r_\epsilon(r_b + r_c) + (1 - \alpha)r_br_c} \\ &+ \frac{(\alpha r_c - 2r_\epsilon)V_2}{r_\epsilon(r_b + r_c) + (1 - \alpha)r_br_c} \\ &+ \left[G_1 + G_4 + \frac{2r_\epsilon + r_b + (1 - \alpha)r_c}{r_\epsilon(r_b + r_c) + (1 - \alpha)r_br_c}\right]V_3. \end{aligned} \tag{12}$$

For the sake of clarity, these equations are rewritten below in a form which places in evidence the terms obtained from the table.

$$\begin{aligned} 0 = Y_{c_1\epsilon_1}E_1 &+ (G_2 + G_3 + Y_{b_2b_2} + Y_{c_1c_1})V_2 \\ &+ (Y_{c_1b_1} + Y_{b_2c_2})V_3 \end{aligned} \tag{13}$$

$$\begin{aligned} 0 = Y_{b_1\epsilon_1}E_1 &+ (Y_{c_2b_2} + Y_{b_1c_1})V_2 \\ &+ (G_1 + G_4 + Y_{c_2c_2} + Y_{b_1b_1})V_3. \end{aligned} \tag{14}$$

It will be observed in this case, as in the preceding example, most of the terms in the admittance coefficients are obtained from the table.

When the circuit configuration is such that the mesh method of analysis is better suited that the node method, one may use the dual of the procedure outlined above in conjunction with appropriate tables of impedance coefficients.[5] In the case of the vacuum tube a difficulty arises in that the impedance coefficients associated with the grid terminal are infinite. One way of circumventing this difficulty is to assume that the tube has a grid return. With this assumption and with the meshes numbered as in Fig. 4(a), the impedance coefficients table assumes the following form

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[5] $Z_{\nu\mu}$ is numerically equal to the voltage drop induced in the νth mesh by a unit current in the μth mesh when all voltage sources are short-circuited and all meshes except the μth mesh are open-circuited. In the case of a passive network which does not contain coupled coils and in which all mesh currents have the same positive sense, $Z_{\nu\nu}$ is equal to the sum of the impedances of the branches traversed by the νth mesh current, and $Z_{\nu\mu}$ $(\nu \neq \mu)$ is equal to the negative of the impedance of the branch common to νth and μth meshes. It should be noted that when the mesh method of analysis is used, the network $N-\Delta$ consists of N with Δ replaced by a short circuit.

TABLE III

Impedance Coefficients Table for the Vacuum Tube with a Grid Return

$Z_{11}=R_g$	$Z_{12}=0$	$Z_{13}=-R_g$
$Z_{21}=\mu R_g$	$Z_{22}=r_p$	$Z_{23}=-\mu R_g-r_p$
$Z_{31}=-(1+\mu)R_g$	$Z_{32}=-r_p$	$Z_{33}=r_p+(1+\mu)R_g$

where R_g is the grid return resistance, r_p is the plate resistance and μ is the voltage amplification factor.

In the case of the transistor there are no such difficulties, and the impedance coefficients (with the meshes numbered as in Fig. 4(b)) read

TABLE IV

Impedance Coefficients Table for the Transistor

$Z_{11}=r_e+r_b$	$Z_{12}=-r_b$	$Z_{13}=-r_e$
$Z_{21}=-r_m-r_b$	$Z_{22}=r_b+r_c$	$Z_{23}=r_m-r_c$
$Z_{31}=r_m-r_e$	$Z_{32}=-r_c$	$Z_{33}=r_e+r_c-r_m$

where the symbols have the same meaning as in Table I, and $r_m=\alpha r_c$.

In conclusion, it will be noted that the method outlined here has a twofold advantage over the conventional method. First, its use does not require a redrawing of the circuit diagram, and second, the process of setting up the node (or mesh) equations is simplified considerably through the use of admittance or impedance coefficients tables. It will also be noted that the method can readily be extended to networks containing active elements other than three-terminal vacuum tubes and transistors, by constructing admittance or impedance coefficients tables for such elements.

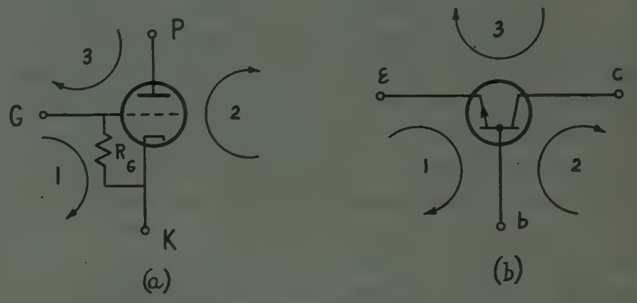

Fig. 4—(a) Vacuum tube with a grid return. (b) Transistor.

Correction

Richard F. Shea, author of the paper, "Transistor Operation: Stabilization of Operating Points," which appeared on pages 1435–1437 of the November, 1952, issue of the Proceedings of the I.R.E., has brought the following errors to the attention of the editors:

On page 1435, Equation (3), $I_j=0$ should read $I_e=0$. Equation (9), R_{DC} should be P_{DC}.

On page 1436, the eleventh line below Fig. 2 should be $P_{DC}=10.8$ mw instead of 14.6 mw; two lines further, 8.6 mw should be 4.8 mw. Equation (14), S in denominator should be removed; third line below Fig. 3, 5 mw should be 1.4 mw.

An Analysis of Magnetic Shift Register Operation*

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Summary—The purpose of this paper is to describe quantitatively the design of a magnetic shift register. The article is divided into 3 sections as follows:

A. A qualitative description of magnetic shift register operation.
B. The design of a magnetic shift register.
C. The effects of leakage inductance on stability and maximum speed of operation of a magnetic shift register.

In section B, the equivalent impedance concept discussed in a previous article by the author[1] is used in the quantitative analysis of a magnetic shift register. Optimum design conditions are derived and variations from optimum design discussed.

In section C, the maximum and minimum allowable values of leakage inductance are found such that the shift register remains stable and information is reliably transferred from core to core.

A. A QUALITATIVE DESCRIPTION OF MAGNETIC SHIFT REGISTER OPERATION

FIG. 1 SHOWS A TYPICAL SECTION of a magnetic shift register consisting of 4 magnetic cores. Information is put into the first core by means of the input driver, and then advanced to the second core by means of the A driver. Fig. 2 shows the relationship in time of the A driver current, the B driver current, and the input driver current. The information being put into the line in Fig. 2 is the binary number one zero. The input driver current is in such a direction as to drive the first core into positive saturation (B_M). The A and B driver currents are in such a direction as to return the cores to a condition of negative saturation ($-B_M$). The core material used has a rectangular hysteresis loop as shown in Fig. 3. When a core in the shift register is in the positive residual flux density condition, it is defined as having a "one" stored in it, and when it is in the negative residual flux, it is defined as having a "zero" stored in it.

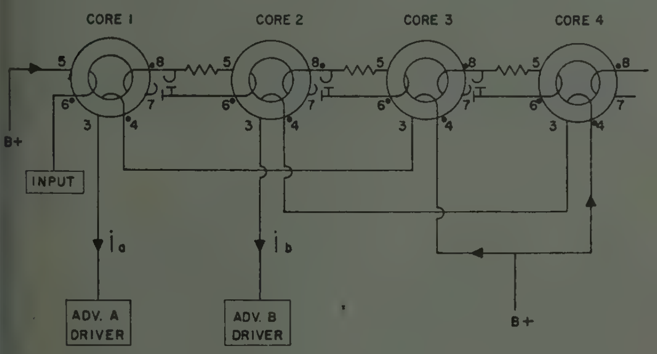

Fig. 1—Four-core magnetic-shift register. Advance winding 3–4. Input winding 5–6. Output winding 7–8.

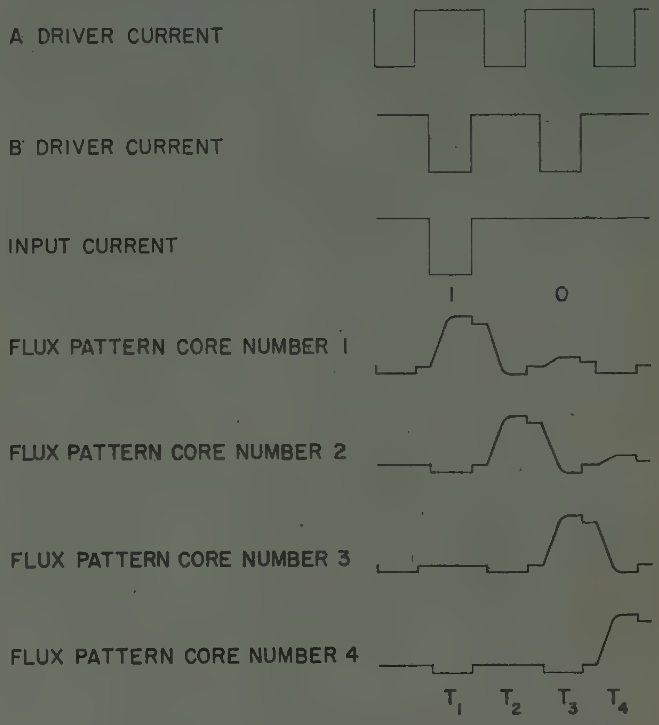
A DRIVER CURRENT
B DRIVER CURRENT
INPUT CURRENT
1
0
FLUX PATTERN CORE NUMBER 1
FLUX PATTERN CORE NUMBER 2
FLUX PATTERN CORE NUMBER 3
FLUX PATTERN CORE NUMBER 4
T_1
T_2
T_3
T_4

Fig. 2—Magnetic-shift register pulse pattern configuration.

The cycle of operation in putting the binary number one zero into the shift register will now be described. Let it be assumed that all four cores are initially in the "zero" state, i.e., they are at the value $-B_R$ in Fig. 3. The input driver and the B driver are pulsed, and produce currents shown in Fig. 2. These are coincident in time. At time T_1 (Fig. 2), the input current flows through the input winding in such a direction as to change the flux in core 1 to the condition B_M. The B-driver current flows through the advance windings of cores 2 and 4 in such a direction as to magnetize them to $-B_M$. The voltage pulse produced at the terminals of the output winding of core 1 as its flux changes from $-B_R$ to B_M is in such a direction as to drive current through the back resistance of the series rectifier in the transfer loop between cores 1 and 2. Therefore, no appreciable demagnetizing current flows into core 2.

At the end of the input and B current pulses, core 1 is at a value of B_R, and all other cores are at a value $-B_R$. At time T_2, the A driver is pulsed. Its current produces a magnetizing force in such a direction as to return cores 1 and 3 to negative saturation. The change in flux in core 1 produces a voltage at its output-winding terminals of polarity to drive current through the forward conduction direction of the series rectifier and into

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[1] E. A. Sands, "The behavior of rectangular hysteresis loop magnetic materials under current pulse conditions," PROC. I.R.E., vol. 40, pp. 1246–50; October, 1952.

core 2. This current is in such a direction as to magnetize core 2 to the condition B_M. At time T_3, the B current pulse changes core 2 to "zero," and advances the information into core 3 in the same manner as described for the transfer from 1 to 2. The voltage appearing across the input terminals (5–6) of core 2 is in such a direction as to drive current into the output winding of core 1 through the forward conduction direction of the series rectifier. This effect, which is called "back flow of information," is manifested in the small positive bump occurring in the flux pattern of core 1 at time T_3, and the flux pattern of core 2 at time T_4.

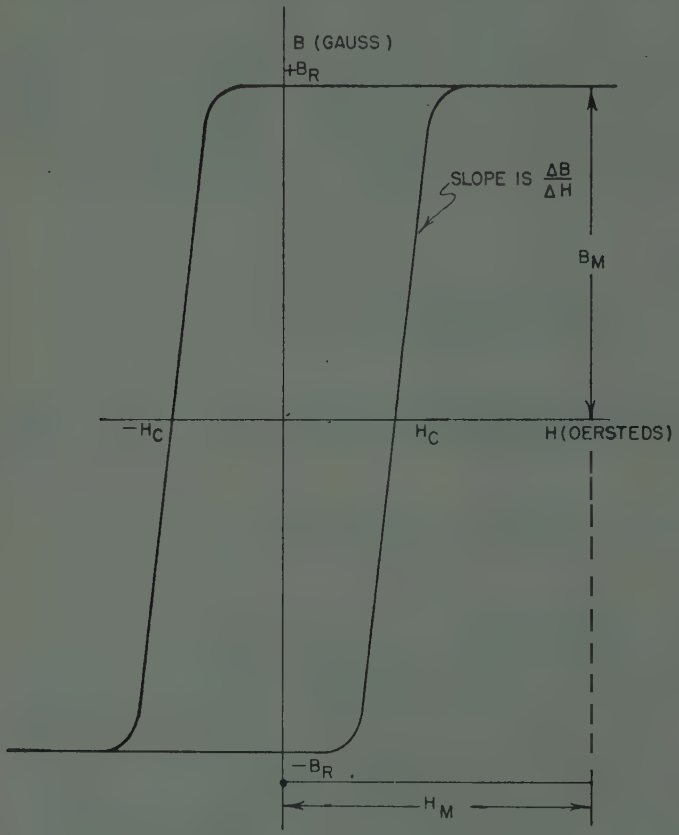


Fig. 3—Pertinent quantities on static B-H loop.

The back flow of information must be less than a certain value to maintain a good zero-to-one signal ratio, and to maintain stability. It is the function of the series resistor and shunt rectifier to cut down the amount of back flow of information. At time T_3, the voltage across the input terminals of core 2 is in such a direction as to make the shunt rectifier conduct. The series resistor is adjusted to cause a sufficient voltage drop across it that core 1 is not appreciably magnetized. At time T_4, the A driver current pulse changes core 3 to the "zero" state, and the information is transferred to core 4.

B. The Design of a Magnetic Shift Register

The design problem to be considered is that of reliably transferring a "one" from one core to another in a given time and using a minimum amount of input energy. Since the magnetic shift register is an iterative structure, and the transfer is concerned with 3 cores, namely the core out of which the "1" is being read and the cores preceding and succeeding this core, for the purposes of analysis, cores 1, 2, and 3 of Fig. 1 will be considered as a typical set of 3 and the situation corresponding to the transfer of a "1" from core 2 to core 3 will be analyzed. In the previously cited article by the author[1] an expression was derived for the equivalent input resistance per turns squared ($\overline{Ro}$) of a rectangular hysteresis loop magnetic material operating on its saturation hysteresis loop under current pulse conditions. This was

$$\overline{Ro} = \frac{A(B_R + B_M) \times 1.26 \times 10^{-8}}{H_M T L} \text{ (ohms/turns squared)}$$

where

B_R = Residual flux density in gauss.
B_M = Saturation flux density produced by H_M in gauss.
H_M = Amplitude of the step function of applied field in oersteds.
A = Cross-sectional area of the core in centimeters squared.
L = Mean path length of the core in centimeters.
T = The time it takes for the core to change its condition from $-B_R$ to B_M with the application of H_M (measured in seconds).

Fig. 4 shows a typical switching time applied field-strength characteristic (H_M vs T) for orthonik cores of 1.1 and 0.5 mil thickness. The product $H_M T$ is approximately constant in the range 1 to 5 oersteds which means that $\overline{Ro}$ is approximately constant and independent of applied field strength in this range.

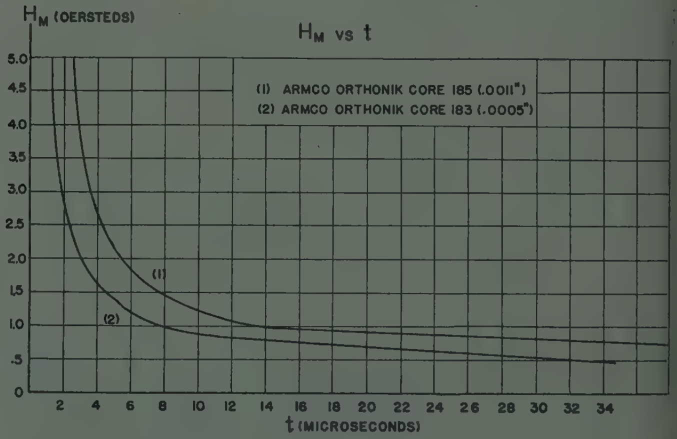


Fig. 4

Fig. 5 is the equivalent circuit corresponding to the condition of the transfer of a "1" from core 2 to core 3 of Fig. 1. N_3, N_5, N_7 correspond respectively to the advance winding (3–4) the input winding (5–6), and the output winding (7–8). The series resistor in the transfer loop between cores is called R_1, and the forward resistance of the series and shunt rectifiers is called R_r

(assumed to be constant). $N_5^2\overline{R}o$ is the equivalent resistance of core 3 as viewed from its input winding. $N_7^2\overline{R}o$ is the equivalent resistance of core 1 as viewed from its output winding. $N_3^2\overline{R}o$ is the equivalent resistance of core 2 as viewed from its advance winding.

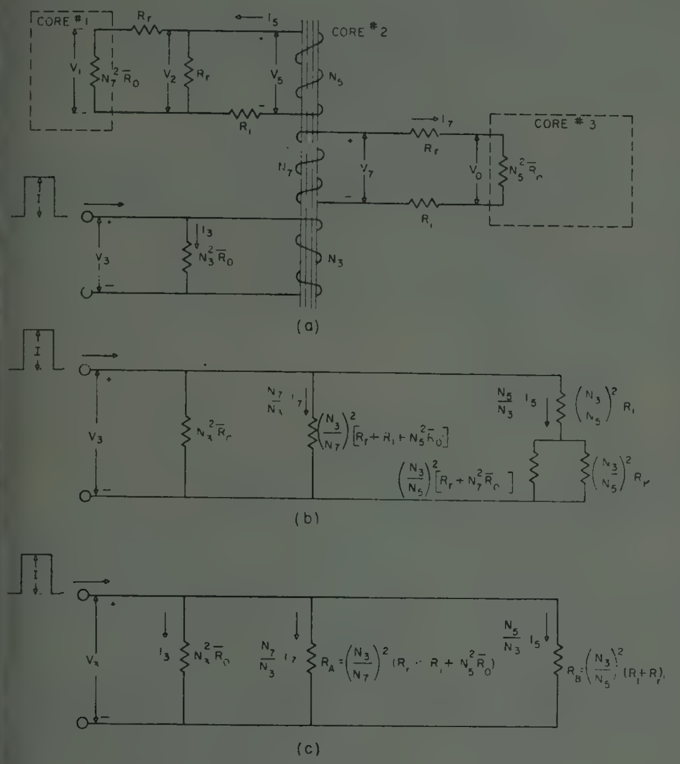

CORE #1
CORE #2
CORE #3
R_r
R_1
I_5
I_7
I_3
V_1
V_2
V_5
V_7
V_0
V_3
$N_7^2\overline{R}_0$
$N_5^2\overline{R}_r$
$N_3^2\overline{R}_0$
N_5
N_7
N_3
I
(a)
$\frac{N_7}{N_3} I_7$
$\frac{N_5}{N_3} I_5$
$\left(\frac{N_3}{N_7}\right)^2[R_r+R_1+N_5^2\overline{R}_0]$
$\left(\frac{N_3}{N_5}\right)^2 R_1$
$\left(\frac{N_3}{N_5}\right)^2[R_r+N_7^2\overline{R}_0]$
$\left(\frac{N_3}{N_5}\right)^2 R_r$
(b)
$R_A=\left(\frac{N_3}{N_7}\right)^2(R_r + R_1+N_5^2\overline{R}_0)$
$R_B=\left(\frac{N_3}{N_5}\right)^2(R_1+R_r)$
(c)

Fig. 5—Equivalent circuit for transfer of a one.

In the analysis, the following additional symbols will be used

T_3 = Time taken by core 3 to change from "0" to "1" during transfer.

T_2 = Time taken by core 2 to change from "1" to "0" during transfer.

F_{ij} = Fractional amount of total possible flux change $((B_R+B_M)A)$ in core i $(i=1, 2, 3)$ at time T_j $(j=2, 3)$. By definition $F_{ij} \leqq 1$.

R_{in} = Equivalent input resistance as seen at the advance winding terminals of core 2 during the time interval $0 \leqq t \leqq T_3$.

$$N = \frac{R_1+Rr}{N_5^2\overline{R}o}$$

$$\left(\frac{d\phi}{dt}\right)_i = \text{Rate of change of flux with time in core } i \qquad (i=1, 2, 3).$$

In what follows, these physical facts must always be borne in mind:

A. In a shift register in which information is being transferred reliably, the flux density change in going from "0" to "1," or from "1" to "0" is the same in all cores and is identically (B_R+B_M).

B. A core out of which a "1" is being transferred (for example core 2) must change from "1" to "0" at a slower rate than the core which it is driving changes from "0" to "1." This follows from the fact that after the driving core saturates, no further energy can be transferred by transformer action to the driven core. Therefore, $T_2 \geqq T_3$.

C. The input to the advance winding is a constant amplitude step-function of current lasting long enough to insure that all flux changes have been completed.

D. The cores are identically wound, have equal cross-sectional areas, and mean path lengths.

The relationship between N_7 (the output winding) and N_5 (the input winding) will now be considered. From Fig. 5(a), the voltage across the output winding of core 2 is

$$V_7 = N_7\left(\frac{d\phi}{dt}\right)_2 \times 10^{-8} = \frac{N_7F_{23}(B_R+B_M)A \times 10^{-8}}{T_3} \qquad 0 \leqq t \leqq T_3. \quad (1)$$

(1) follows from the fact that the equivalent circuit is entirely resistive and the input is a constant amplitude pulse of current. Similarly, the voltage across the input winding of core 3 is

$$Vo = \frac{N_5(B_R+B_M)A \times 10^{-8}}{T_3} \qquad 0 \leqq t \leqq T_3. \quad (2)$$

From the equivalent circuit

$$Vo = \frac{V_7N_5^2\overline{R}o}{R_1+Rr+N_5^2\overline{R}o} = V_7/1+N. \quad (3)$$

Equating (2) and (3) and substituting (1) in (3), the relationship between the number of turns on the output and input windings is found to be

$$N_5 = F_{23}N_7/1+N. \quad (4)$$

Therefore, the output winding always has to have more turns than the input winding.

The flux change in core 1, when a "1" is being transferred from core 2 to core 3 will now be found. This is the back flow of information mentioned previously. For the equivalent representation of Fig. 5(a), the back flow of information is independent of T_2 and T_3, so for simplicity, the case where $F_{23}=1$ will be discussed (core 2 and core 3 switch in the same time), and $F_{12}=F_{13}$ found. This is the "0" to "1" ratio and should be less than 1 to 8. The total flux change in core 1 is

$$F_{12}(B_R+B_M)A = \frac{1}{N_7}\int_0^{T_2} V_1 dt \times 10^{+8}. \quad (5)$$

From Fig. 5(a):

$$V_1 = \frac{V_5(N_7^2\overline{R}oRr)}{R_1(2Rr+N_7^2\overline{R}o)+Rr(N_7^2\overline{R}o+Rr)}. \quad (6)$$

In the usual case, $N_7^2\overline{Ro}$ is much greater than Rr. Making use of this in (6)

$$V_1 = V_5\left(\frac{Rr}{R_1 + Rr}\right), \tag{7}$$

but

$$V_5 = \frac{N_5(B_R + B_M)A \times 10^{-8}}{T_2}. \tag{8}$$

Substituting (8) and (7) in (5)

$$F_{12} = \frac{N_5}{N_7}\,\frac{Rr}{R_1 + Rr}. \tag{9}$$

There are then, two ways to make the "0" to "1" ratio small. First, R_1 can be made large with respect to Rr. Second, the number of turns on the output winding can be made great—relative to the number of turns on the input winding. However, the turns ratio is determined from minimum input energy considerations, and, as will be shown, stability obtained by increasing N_7/N_5 can only be obtained by increasing the input ampere turns N_3I.

An optimum condition will now be found such that N_5I_7 is a maximum for a given input N_3I. This means that a maximum number of ampere turns is applied to core 3 for a given number of ampere turns applied to the advance winding of core 2. It defines the minimum possible input energy to transfer a "1" in a given time. It will be assumed that $N_7^2(\overline{Ro})$ is much greater than Rr so that the equivalent circuit of Fig. 5(b) reduces to that of 5(c). From 5(c)

$$I_7 = \frac{N_3IN_7\overline{Ro}(R_1 + Rr)}{(Rr + R_1 + N_5^2\overline{Ro})^2 + N_7^2\overline{Ro}(R_1 + Rr)}. \tag{10}$$

In this expression I, $\overline{Ro}$, N_3, R_1, and Rr are constant. N_5 will be held constant, and a value of N_7 found which will make I_7 a maximum, thereby making N_5I_7 a maximum. Setting $dI_7/dN_7 = 0$, and solving

$$\frac{N_7}{N_5} = \frac{1 + (F_{23})^2}{F_{23}} \quad \text{and} \quad N_5I_7 = N_3I/2\left(1 + \frac{1}{F_{23}}\right). \tag{11}$$

Examination of (11) shows that N_5I_7 has its maximum possible value for $F_{23} = 1$. Corresponding to this value of F_{23}

$$\frac{N_7}{N_5} = 2; \quad R_1 + Rr = N_5^2\overline{Ro}; \quad 4N_5I_7 = N_3I. \tag{12}$$

Physically, these results mean that in an optimum design, the flux change in cores 2 and 3 proceed at the same rate ($T_2 = T_3$), and the ampere turns applied to the advance winding of core 2 equals 4 times the ampere turns applied to the input winding of core 3. To achieve an optimum design two things must be done. The turns ratio must be made 2 to 1, and the sum of the rectifier resistance (Rr) and the series resistance (R_1) must be made equal to the equivalent resistance of the input winding of core 3. Before discussing variations from optimum conditions, an example will be worked to make the design procedure clear.

The cores to be used are made up of 2 wraps of $\frac{1}{8}$ inch wide, 1.1 mil thick orthonik. The core has a mean-path length of 1.2 inches. Switching time $T_2 = T_3 = 8$ μsec. The rectifier selected has a forward resistance = 100 ohms. The "0" to "1" ratio = $\frac{1}{8}$. From this

$H_M = 1.5$ oersteds (see Fig. 4 H_M vs T curve).
$N_5I_7 \doteq 2H_ML = 3.6$ ampere turns.
$\overline{Ro} = 1.85 \times 10^{-2}$ ohms per turns squared.
$R_1 = 300$ ohms (equation (9), $N_5/N_7 = .5$, $F_{12} = F_{23} = 1$).
$N_5^2\overline{Ro} = 400$ ohms (equation 12).

Using these values

$N_5 = 147$ turns
$N_7 = 294$ turns
$N_3I = 14.4$ ampere turns

The choice of N_3 depends on the current available from the advance driver tube.

The optimum design condition can be equivalently stated as follows: For a given value of N_3I, the fastest possible transfer time is realized when $N_7/N_5 = 2$, and $R_1 + Rr = N_5^2\overline{Ro}$. In a practical design, however, $R_1 + Rr$ should be chosen to be smaller than $N_5^2\overline{Ro}$ to allow for the variation of $\overline{Ro}$ from core to core, to allow for the effects of leakage inductance and rectifier capacitance. The variation of the transfer time (T_2) as the ratio $R_1 + Rr/N_5^2\overline{Ro} = N$ is varied will now be considered.

The circuit of Fig. 5(c) reduces to that of Fig. 6(a). As N is made smaller core 2 takes an increasingly longer time to switch than core 3. At time T_3, when core 3 saturates, its input impedance changes from $N_5^2\overline{Ro}$ to zero. Therefore, the equivalent circuit of 6(a) must be replaced by that of 6(b) after time T_3, and is valid until time T_2, when core 2 saturates. Switching times T_2 and T_3 will now be found as functions of N. The total equivalent input resistance as seen at the advance winding terminals (N_3) of Fig. 6(a) is

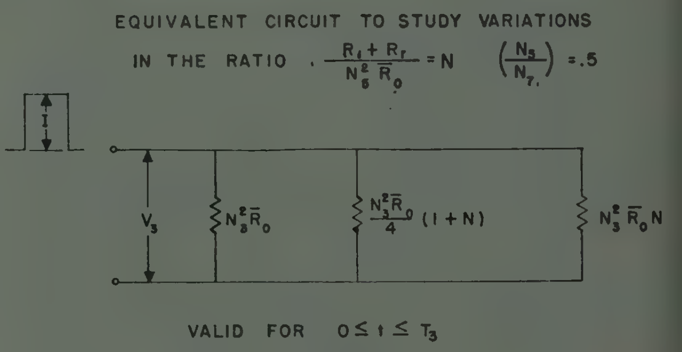

EQUIVALENT CIRCUIT TO STUDY VARIATIONS
IN THE RATIO . $\frac{R_1 + R_r}{N_5^2\overline{R}_0} = N$ $\left(\frac{N_5}{N_7}\right) = .5$
I
V_3
$N_3^2\overline{R}_0$
$\frac{N_3^2\overline{R}_0}{4}(1+N)$
$N_3^2\overline{R}_0N$
VALID FOR $0 \leq t \leq T_3$

Fig. 6(a)

$$(R_{in})_3 = \frac{N_3^2\overline{Ro}(1 + N)(N)}{N^2 + 6N + 1}. \tag{13}$$

At time T_3, from the voltage-flux time relations in core 3

$$\frac{(B_R + B_M)A}{10^8} = \frac{2IR_{in}T_3}{N_3(1+N)}, \tag{14}$$

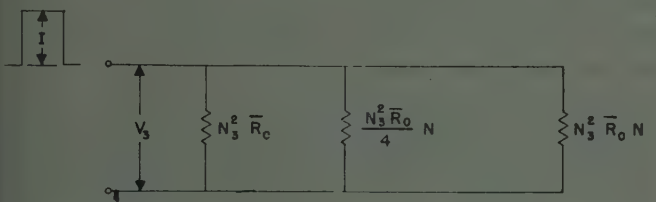
so that

$$T_3 = \frac{(B_R + B_M)A(N^2 + 6N + 1) \times 10^{-8}}{2N_3I\overline{R}oN}. \tag{15}$$

The flux change in core 2 at time T_3 is

$$F_{23}(B_R + B_M)A = \frac{1+N}{2}(B_R + B_M)A \quad \text{from (4).} \tag{16}$$

T_3 is a minimum for $N=1$, which confirms the fact that the fastest possible transfer time corresponds to $R_1+Rr = N_5^2\overline{R}o$, and $N_7/N_5=2$.


I
V_3
$N_3^2\ \overline{R}_c$
$\frac{N_3^2\overline{R}_0}{4}\ N$
$N_3^2\ \overline{R}_0\ N$

VALID FOR $T_3 < t \leq T_2$

Fig. 6(b)

After core 3 has switched, core 2 works into the circuit of Fig. 6(b) for which the equivalent input resistance is

$$(R_{in})_2 = \frac{N_3^2\overline{R}oN^2}{5N + N^2}. \tag{17}$$

To find T_2, use is made of the fact that the ratio of V_3 before core 3 has switched to V_3 after core 3 has switched is the same as the ratio of $(R_{in})_3/(R_{in})_2$; so that

$$\frac{F_{23}/T_3}{\dfrac{1-F_{23}}{T_2-T_3}} = (R_{in})_3/(R_{in})_2 = \frac{(N+1)(N+5)}{N^2+6N+1}. \tag{18}$$

Substituting (15) and (16) in (18) and solving for T_2

$$T_2 = T_3\frac{(2N+6)}{N^2+6N+1}$$
$$= \frac{(B_R + B_M)A(2N+6) \times 10^{-8}}{(2N_3I\overline{R}o)(N)}. \tag{19}$$

The table below gives T_2/T_3 and

$$T_2\left(\frac{2N_3I\overline{R}o \times 10^8}{(B_R + B_M)A}\right) = D$$

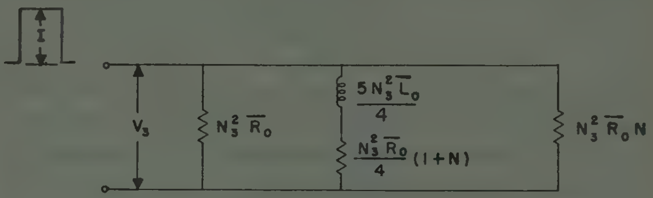
as functions of N

N	D	T_2/T_3
1	8	1
0.8	9.5	1.18
0.6	12	1.45
0.5	14	1.65
0.3	22	2.29

SECTION C. THE EFFECTS OF LEAKAGE INDUCTANCE ON STABILITY AND MAXIMUM SPEED OF OPERATION OF A MAGNETIC SHIFT REGISTER

Of great importance in the design of a magnetic shift register is the number of cores that can be driven from a given sized driver tube. To reduce the energy dissipated per core, it is desirable to reduce the cross-sectional area of the iron to a minimum. The lower limit on the amount of iron to be used is the maximum allowable leakage inductance in the transfer loop. If there is too much leakage inductance, a "1" cannot be transferred reliably.

Fig. 6(c) shows an equivalent circuit which takes into account the leakage inductance in the output winding of core 2, and the input winding of core 3. All quantities are referred to the input winding N_3. $\overline{L}o$ is the leakage inductance per turns squared. It can be shown that


I
V_3
$N_3^2\ \overline{R}_0$
$\frac{5N_3^2\overline{L}_0}{4}$
$\frac{N_3^2\overline{R}_0}{4}(1+N)$
$N_3^2\ \overline{R}_0\ N$

EQUIVALENT CIRCUIT TAKING INTO ACCOUNT LEAKAGE INDUCTANCE

Fig. 6(c)

$$\frac{1}{F_{23}} = \frac{2(1+N)\left\{1 - \left(\dfrac{T}{T_3}\right)(1 - e^{-T_3/T})\right\}}{(1+N)^2 + 4N\dfrac{T}{T_3}(1 - e^{-(T_3/T)})}$$

$$\text{for } N_7/N_5 = 2, \tag{20}$$

where

$$T = \frac{5\overline{L}o}{\overline{R}o} \times \frac{(1+N)}{(1+6N+N^2)}.$$

The condition for maximum allowable leakage inductance consistent with a reliable transfer is found by setting $F_{23}=1$. For all values of F_{23} greater than 1, core 2 switches before core 3. Fig. 7 is a plot of the maximum allowable N ($N_{\max}$), such that $F_{23}=1$, for various values of T/T_3. For any value of N greater than $N_{\max}$, there will not be a reliable transfer. On the same curve is plotted the ratio as a function of T/T_3

$$\frac{T}{\overline{L}o/\overline{R}o}.$$

The way these curves may be used will be illustrated in the following example for which $N=0.81$ and $T_3=8$ μsec. From Fig. 7 (page 998) for this value of N

$$\frac{T}{T_3} = 0.05, \text{ so that } T = 0.4\ \mu\text{sec.}$$

and

$$\frac{\overline{Lo}}{\overline{Ro}} = \frac{T}{1.37}.$$

Using the core in the example in Section B,

$$\overline{Ro} = 1.85 \times 10^{-2} \text{ ohms per turns squared,}$$

so that

$$\overline{Lo} = 0.54 \times 10^{-8} \text{ henries per turns squared.}$$

Therefore, the maximum allowable leakage inductance in the output winding of the previous example is 466 μh if core 3 is to switch in 8 μsec. The expression for $\overline{Lo}$, in this example, can be rearranged as follows

$$\overline{Lo} = \frac{0.05\overline{Ro}T_3}{1.37}.$$

In this form, it is seen that the maximum allowable leakage inductance is proportional to T_3, the switching time of core 3. Therefore, the leakage inductance sets a lower limit on the switching time obtainable with a given core.

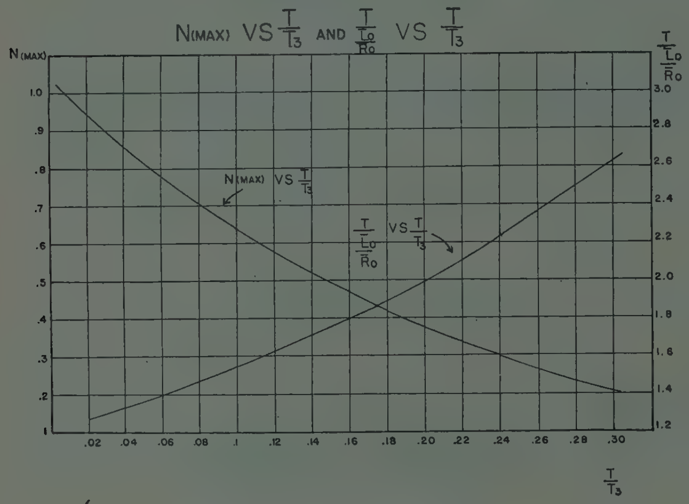


Fig. 7

One of the design criteria previously discussed was the amount of allowable back flow of information. It was stated that the flux change in core 1, when a "1" was being transferred from core 2 to core 3, should not be greater than $\frac{1}{8}(B_R+B_M)A$. Let us assume a flux change of $\frac{1}{8}(B_R+B_M)A$ has taken place in core 1. On the next advance A current pulse, a voltage produced by this back flux appears on the output winding (N_7) of core 1 in such a direction as to magnetize core 2 in the positive flux direction. If the flux change in core 2 is greater than in core 1, the magnetic shift register will become unstable because the noise level builds up in succeeding cores in the same manner as between cores 1 and 2. If the noise flux in core 1 is called $(d\phi)_{1p}$, and the noise flux change produced in core 2 by $(d\phi)_{1p}$ is called $(d\phi)_{2p}$, the necessary condition for stability in the magnetic shift register is that $(d\phi)_{2p}/(d\phi)_{1p}$ be less than 1.

The noise voltage appearing at the output terminals of core 1 lasts about one eighth as long as the voltage pulse occurring when a "1" is read out because the flux change is one eighth as great. Examination of Fig. 7 shows that $N_{\max}$ decreases with T_3. If $\overline{Lo}$ is chosen so that it is about $\frac{2}{3}$ of the maximum allowable leakage inductance per turns squared for the reliable transfer of a "1" in time T_3, then it will be great enough so that for a noise pulse of time duration $T_3/8$, $(d\phi)_{2p}/(d\phi)_{1p}$ is less than 1. If $\overline{Lo}$ is too small, the shift register will become unstable. Therefore, there is an upper limit and a lower limit as well. If the leakage inductance is too large, a "1" cannot be transferred reliably; if it is too small, the shift register becomes unstable. In the example worked out, the maximum value of $\overline{Lo}$ consistent with the reliable transfer of a "1" was $\overline{Lo}=.54\times10^{-8}$ henries per turns squared. Values of $\overline{Lo}$ between 0.067 and 0.54×10^{-8} henries per turns squared would be such as keep the shift register stable and allow for a reliable transfer. Since the maximum allowable $\overline{Lo}$ is proportional to switching time, and the switching time is proportional to total flux change, the larger the noise flux, the less stable the register. For this reason, the stability of the magnetic-shift register is determined by the smallest noise flux such that $(d\phi)_{2p}/(d\phi)_{1p}=1$. Once a noise flux occurs which makes this ratio greater than 1, stability cannot be achieved at some further point in the shift register.

CONCLUSIONS

1. In a magnetic shift register made of identical cores, the minimum possible applied field strength necessary to transfer a "1" from one core to another in a given time T, is $4H_M$.

2. The minimum possible applied field-strength condition is achieved by making the turns ratio of output winding to input winding 2:1, and making the resistance in the transfer loop (R_1+Rr) equal to the equivalent resistance of the core as seen from its input winding, $(N_5^2\overline{Ro})$.

3. The minimum applied field strength condition is not critical with respect to turns ratio, provided the resistance in the transfer loop is properly adjusted. For instance, if a turns ratio of 3 to 1 is selected, and the resistance in the transfer loop is made twice that of the equivalent resistance of the core as viewed from its input winding, the applied field strength need only be increased from $4H_M$ to $4.25H_M$.

4. The applied field strength required to transfer a "1" in a given time T, varies appreciably with reduction of the resistance in the transfer loop from its optimum value. For instance, a reduction of 50 per cent in this resistance from optimum necessitates an increase in applied field strength from $4H_M$ to $6.6H_M$.

5. The maximum allowable leakage inductance for which reliable transfers can be obtained is directly proportional to the product of the equivalent resistance per turns squared and to the switching time.

6. The leakage inductance is primarily responsible for the prevention of noise build-up in a magnetic-shift register.

7. The allowable value of leakage inductance is bounded. It must be great enough to prevent noise build-up, but small enough to allow reliable transfer of information.

ACKNOWLEDGMENTS

The experimental work done was made considerably easier by the excellent pulse control equipment designed by Harold Kenosian and Joseph Chedaker of the Burroughs Adding Machine Company. The author wishes to thank Lyman Orr, Lyle Thompson, Isaac Auerbach, and Felix Kroll of the Burroughs Adding Machine Company for their help, and Martin Littman and Don Dieterle of the Armco Steel Company for their kindness in providing him with magnetic materials and pertinent data.

Debunching of Electron Beams Constrained by Strong Magnetic Fields*

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Summary—**To obtain a clearer insight into the design of velocity modulated tubes, a more complete understanding of the bunching process is required. A theory, to be useful in the design of high power tubes, should include the effects of space charge and large signals. Although such a theory has not yet been developed, a small signal theory, due to Feenberg, including the effects of space charge, has been obtained for some useful physical configurations. This paper deals with the experimental confirmation of this theory and the determination of its range of validity. Numerous experiments were devised to check the various aspects of the theory, and results of these are presented. Tests were made on a conventional two-cavity klystron with a cylindrical drift-tube and gridless gaps at the cavities. Transverse motion of the electrons was constrained by a large axial magnetic field. The extrapolation of the theory beyond its obvious range of validity is shown to be accurate even for large values of the bunching parameter such as to include the first peak of the output current.**

INTRODUCTION

THE FIRST ORDER kinematic theory due to Webster[1] predicts, for a two-cavity klystron,

$$I_2 = 2I_0 J_1(x) \tag{1}$$

where I_2 is the output radio frequency current, I_0 is the direct beam current, and

$$x = \pi N \alpha \tag{2}$$

where N = cycles of transit time in drift-tube and

$$\alpha = \frac{\text{RF voltage}}{\text{DC voltage}}.$$

This expression, however, does not include the effect of the mutual repulsion of electrons due to space charge. In an attempt to include space charge, Webster[2] solved the equations of motion for the infinite beam, assuming positive ion neutralization, and considering longitudinal debunching only (no transverse motion of electrons allowed). The resulting expression for the fundamental component of the output current is given by

$$I_2 = 2I_0 J_1\left(x\,\frac{\sin hl}{hl}\right), \tag{3}$$

provided

$$\left(x\,\frac{\sin hl}{hl}\right)$$

is less than unity. The debunching wave number h is defined by the relation

$$ha = \left(\frac{60 I_0 c}{V_0 u_0}\right)^{1/2} = 174\sqrt{\frac{I_0\ \text{(amps)}}{V_0^{3/2}\ \text{(volts)}}} = 174\sqrt{\text{perveance}} \tag{4}$$

where a is the beam radius, l is the drift-tube length, and u_0 the dc beam velocity. Space charge is seen to manifest itself in a change of the effective bunching parameter. A larger voltage at the input gap is required to maximize the output, but the magnitude of the output current is not affected. The over-all effect on the output radio frequency current is described by the debunching parameter

$$hl = \omega_p T, \tag{5}$$

where $\omega_p = \sqrt{e\rho/m\epsilon_0}$ = plasma frequency and T = dc transit time.

The physical phenomenon that is described by (3) is somewhat as follows. If in a region having uniform density of positive and negative charge (including the case in which the negative charge is moving with uniform velocity), a small region of charge is displaced from its equilibrium position, the restoring forces will cause

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[1] D. L. Webster, "The theory of klystron oscillations," *Jour of Applied Physics*, vol. 10, pp. 864–872; December, 1939.

[2] D. L. Webster, "Cathode-ray bunching," *Jour. of Applied Physics*, vol. 10, pp. 501–508; July, 1939.

this charge to perform an oscillation around this position at the plasma frequency. For the infinite beam, this plasma frequency is a constant depending only on the charge density and, of course, the mass of the particles. The amplitude of the oscillation is determined by the velocity modulation of the electron as it enters the drift tube, since this determines the initial velocity away from its equilibrium position in the beam. The behavior of the electrons is analogous to that of a string of pendulums tied to a belt moving at the beam velocity past a gap which excites oscillations in the pendulum, the initial excitation of each depending on when it passes the gap. Although this description of debunching is satisfactory from the point of view of visualizing the motion of each electron, it turns out that mathematically these oscillations can be described more conveniently in terms of electromagnetic waves propagating through the beam. It can be shown that the fields which cause the above motion of electrons can be expressed as a sum of four waves, each of a different propagation constant γ_n. Since two values of γ_n depend on the charge density in the beam, these waves are known as space charge waves. Their velocities differ only slightly from the beam velocity and they interfere in such a way as to produce the net effect described by (3). The remaining two waves are unaffected by the beam.

A more useful result is that due to Feenberg[3] who has solved the field equations and the force equation subject to the proper boundary conditions for some actual beam configurations. The results are restricted to small signals, nonrelativistic velocities, positive ion neutralization, and non-crossing of electron trajectories. For the finite beam the fields in the drift tube are made up of space charge waves whose propagation constants depend on the transverse boundary conditions as well as the charge density. Moreover, to satisfy the initial conditions at the gaps, infinite sums rather than single terms of such waves must be taken to represent the forced modulation. As an example, for a cylindrical drift tube with no angular variations, the transverse dependence must be of the form $J_0(T_m r)$. Allowed values of T_m are determined by boundary conditions at the walls, a different value of the propagation constant γ_{nm} corresponding to each value of T_m. In order to satisfy the conditions at the gaps, the known modulation voltage at the input must be expressed in an infinite sum of $J_0(T_m r)$, hence the total fields are of the form

$$\sum_{n=1}^{4}\sum_{m=1}^{\infty}(A_{mn})e^{\gamma_{nm}z}J_0(T_m r) \qquad (6)$$

The physical gap configuration determines how many terms of this infinite expansion in "m" are important. Strictly speaking, each term of this infinite series corresponds to space charge waves of different plasma oscillation frequency, which is no longer determined by the charge density alone, but by the finiteness of the drift tube as well. The fields in the drift tube can thus be considered as a superposition of many space charge waves of different plasma frequencies. With a large axial magnetic field two of these infinite sums drop out, and the resulting output current can be written as

$$I_2 = \frac{I_0 x F\left(hl,\, r,\, k'a,\, \frac{b}{a}\right)}{l} \qquad (7)$$

where F/l is a function tabulated by Feenberg (Fig. 1), r is the radial position, $k'=\omega/u_0$, u_0 is the dc beam velocity, a is the beam radius, and b is the drift tube radius. Debunching in the finite beam is thus expressed by the function F, given by Feenberg and reproduced in Fig. 1 for a typical condition. As before, it is a function of the debunching parameter hl, but in addition depends on the radial position r and the normalized beam and drift tube radii $k'a$ and $k'b$.

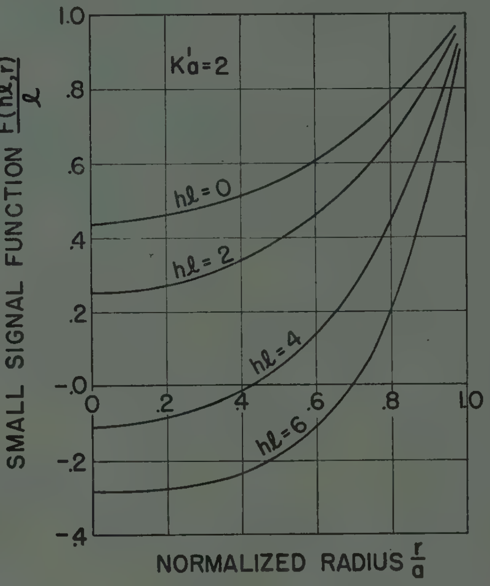

Fig. 1—Feenberg's small signal function for $k'a=2$ for gridless gaps, infinite magnetic field, and beam filling cylindrical drift tube.

Feenberg's theory may be arbitrarily extrapolated to include large values of the bunching parameter by considering both the exact solution of the small signal problem and Webster's first order bunching theory. For small values of x, (1) reduces to

$$I_2 = I_0 x. \qquad (8)$$

On the other hand, the small signal solution of (7) implies that the effective bunching parameter now has a value of $(x(F/l))$. Using this value of the bunching parameter, as obtained by Feenberg for the space charge problem, the output current for large bunching parameters can be written in terms of the usual Bessel function without regard for the fact that the effective bunching parameter was calculated on the basis of small signals. Thus

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[3] E. Feenberg, "Notes on velocity modulation," Sperry Gyroscope Company Report No. 5221-1043; 1945.

$$I_2 = 2I_0J_1\left\{x\,\frac{F(hl,\,r,\,k'a,\,b/a)}{l}\right\} \tag{9}$$

which for small signals reduces to the proper form. The validity of such an arbitrary extrapolation can only be checked experimentally and is confirmed by the experiments described later in the paper.

General Remarks

An inspection of (9) and the Feenberg function F reveals some interesting features of the output current.

(a) Since the small signal function F depends on radial position, the output current density is no longer constant across the beam. In general, the current density near the center is smaller, since debunching forces are stronger there.

(b) The rf current density decreases for increasing values of hl but not in a simple way, as in the case of the infinite beam.

(c) For a sufficiently large value of hl, the output current density becomes zero at some radial position regardless of the value of the bunching parameter (i.e., driving voltage), since the function F itself becomes zero (Fig. 1). This first occurs at $r=0$ and moves outward as hl is increased. The existence and location of this critical radius would lend strong support to the theory.

(d) For still larger values of hl, the output current density near the center actually becomes negative. This reversal should be experimentally observable. Mathematically, the reason for this reversal of current density can be seen from the nature of the solution required by presence of walls surrounding the beam. As stated previously, to satisfy the initial conditions at the gap, it is necessary to superimpose an infinite set of waves with different propagation constants. The motion of the beam is described by adding an infinite set of plasma oscillations of different amplitudes with slightly different frequencies. Since each plasma wave has a different radial dependence, their sums will interfere differently at various radii; in particular, near the center of the beam these waves interfere in such a way that electrons there will reach their maximum displacement earlier than those at the edge, start moving back toward equilibrium, pass through it, and have displacements from equilibrium of the opposite sign. This results in a current density at the center of opposite sign to that at the edge, and indeed, one can make the obvious statement that electrons right at the edge of the wall must have their plasma oscillations superimposed in such a way that there is no debunching at all. Thus the electrons near the wall never have any oscillatory behavior, since the longitudinal field at the wall must be zero.

Alternately, this may be seen by considering the debunching forces themselves. At the wall these forces are neutralized by image forces outside the beam, but at the center, these image forces are not as effective, resulting in stronger debunching. For sufficiently high current densities (large values of hl), the electron charge density may be such as to cause an anti-bunch to be formed at the center, while at the wall, a bunch may exist.

Description of Experiments

(a) *Tube Design.* To investigate the various aspects of (9) described in the previous section, a conventional two-cavity klystron was constructed. A strong axial magnetic field was used to constrain the beam. Eq. (9) shows that it is sufficient to specify the parameters hl, $k'a$, b/a, and the value of x to completely determine the normalized output current. The tube was designed to include a sufficiently wide range of these parameters to be useful in tube design and at the same time provide adequate information about regions where debunching forces are large. The parameter hl was varied by varying the tube perveance (changing cathode-anode spacing), the parameter $k'a$ by changing the dc beam voltage, and the value of x by changing the input power to the bunching cavity.

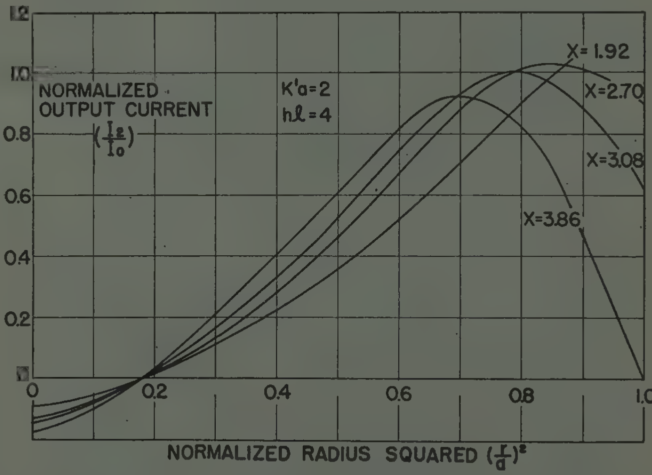

Fig. 2—Typical contributions to the output current from various radial positions for several values of the bunching parameter.

(b) *Interpretation of* (9). Eq. (9) gives two kinds of results: first, the current density as a function of radius, and second the total current which can be obtained by averaging these contributions over the beam cross-section. A typical variation of the normalized output current as a function of radius for $k'a=2$ and $hl=4$ is shown in Fig. 2. The total current (Fig. 3) is proportional to the area under this curve for each value of the bunching parameter. Universal curves for $k'a=2$ and 3 are shown in Figs. 4 and 5. These curves apply to gridless cavities, but can be applied to gridded cavities if the correct function F is taken from reference 3.

(c) *Experiments.* The object of the following experiments was to check the various aspects of these curves

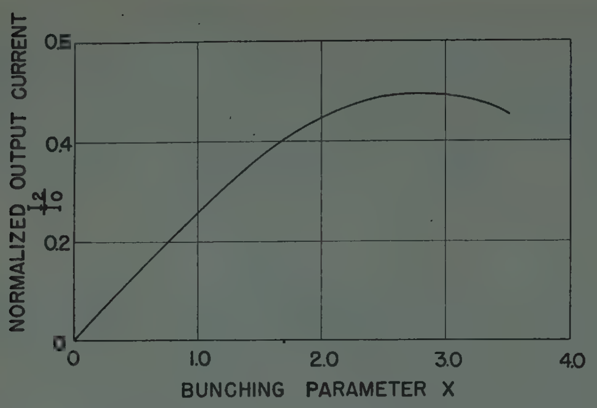

Fig. 3—Output current as a function of bunching parameter.

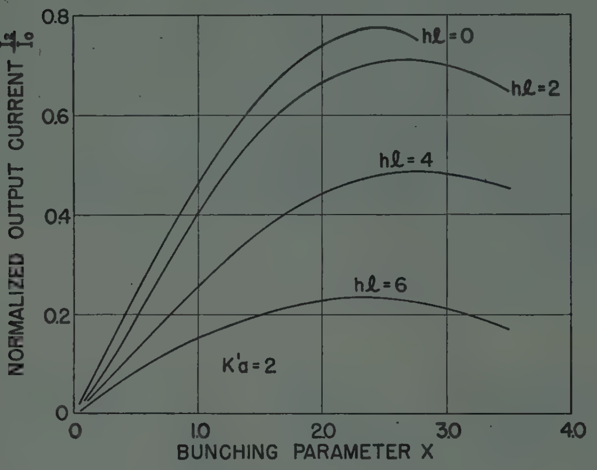

Fig. 4—Design chart for $k'a=2$.

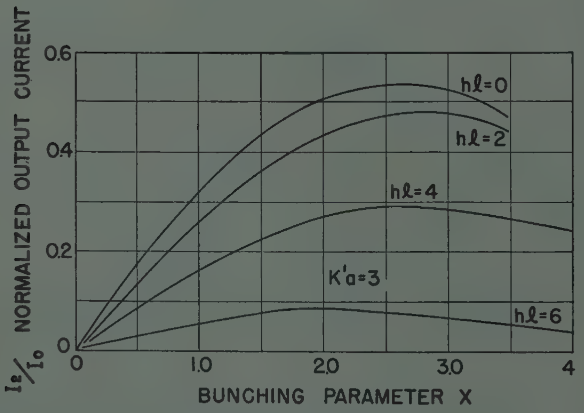

Fig. 5—Design chart for $k'a=3$.

and to see over what range of parameters they apply. The total rf current was measured by sweeping the input power to the buncher cavity dynamically and observing the rectified crystal current from the output cavity on an oscilloscope. Fig. 6 shows the circuit used to make these measurements. To obtain absolute values of the radio frequency current and bunching parameter x, a careful calibration of the microwave equipment was made.

The ratio of shunt impedance R to Q of the cavity is a quantity independent of loss and can be measured by perturbation methods.[4] The Q under operating conditions can be measured by plotting cavity impedance as a function of frequency in a conventional way. The relative variations of rf power at both input and output were obtained using a carefully calibrated square law crystal. Absolute values of power for a given condition were obtained by using a matched bolometer and bridge circuit. Knowing R/Q, Q and the power, the rf voltage across the gap can be calculated. The rf current at the output can be deduced from $I=V/R$, provided the output voltage is sufficiently small compared to the beam voltage. This was achieved by loading down the output cavity.

The individual radial contributions to the output current were measured by placing a thin metallic sheet just ahead of the output cavity. A hole, small compared to the beam diameter, was drilled in this sheet. By means of a bellows arrangement, this sheet, together with the hole, was made to traverse the beam so that the hole explored all radial positions along a given diameter. The small fraction of the rf current reaching the output cavity through this hole was observed dynamically as described. Additionally, the exploring hole was used to check the uniformity of the dc beam in the drift tube. The expected reversal of phase of the output current at a critical radius was checked by comparing its phase with a reference signal from the input, using a standing wave detector padded with matched attenuators at both ends. If a single frequency signal is fed in from one end, no reflections are observed. The same argument is valid for a signal fed in from the other direction. If both signals are fed in simultaneously, a combination of the two waves traveling in opposite directions results in a standing wave. If the phase of one of these is shifted, the position of the node of the standing wave is also shifted; in fact, a phase shift of 180 degrees results in a shift of the node of quarter wavelength. Thus feeding a portion of the input signal of the tube to one end of a

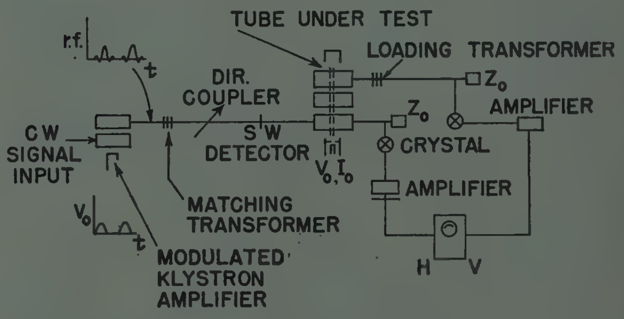

Fig. 6—Experimental setup.

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[4] W. W. Hansen and R. F. Post, "On the measurement of cavity impedance," *Jour. of Applied Physics*, vol. 19, no. 11, pp. 1059–1061; November, 1948.

padded standing wave detector, and the signal from the output cavity to the other end of this detector, a shift of the node should be observed as the exploring hole is moved through the critical radius. This 180 degree phase shift was actually observed. Additional amplification was required to raise the output power level sufficiently, but this only introduces a constant phase shift.

Discussion of Results

The results of these experiments are shown in Figs. 7 and 8 for the case of optimum bunching. It is sufficient to discuss this case since the measured output current as a function of the bunching parameter closely follows the Bessel function $J_1(x)$ (Fig. 3). Before making a comparison with predicted results, the theoretical curves must be modified to include the effect of imperfect beam shape. The theoretical curves assume a completely uniform beam filling the drift tube which was not achieved in practice. However, a correction can be made, since the actual beam distribution was known from exploring hole measurements. The range of modified theoretical values due to this condition is shown in Figs. 7 and 8 together with the experimental values within the accuracy of calibration of the microwave equipment.

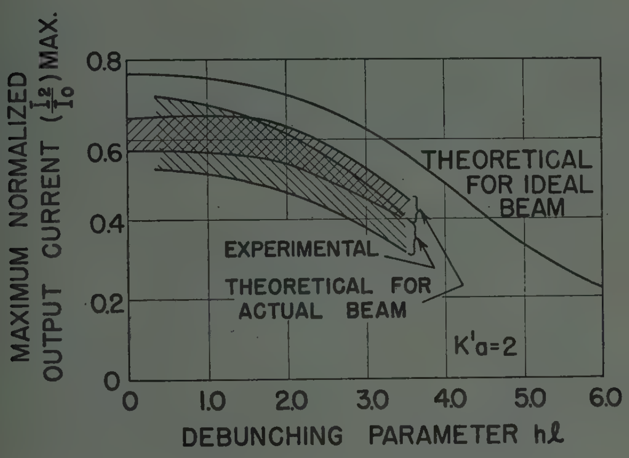

Fig. 7—Comparison of experimental and theoretical results for $k'a=2$.

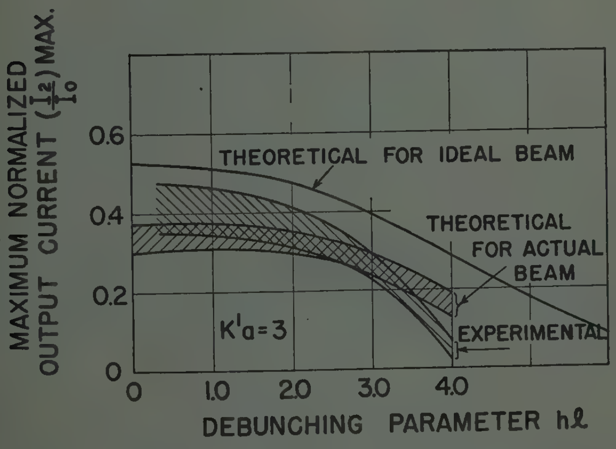

Fig. 8—Comparison of experimental and theoretical results for $k'a=3$.

From these curves, a comparison between theory and experiment can be made. The over-all results of this comparison are:

(a) Small signal space charge theory due to Feenberg is valid if the debunching parameter *hl* is less than 3.5 (i.e., for most practical designs).

(b) The extrapolated theory holds for the same values of *hl* but includes all values of the bunching parameter up to 4 (i.e., well beyond the first peak of output current).

(c) For large values of *hl*, debunching forces appear larger than given by the extrapolated theory.

(d) For large values of the bunching parameter (above 4), experimental results no longer agree with predictions of extended theory. Additional verification was obtained from cross-sectional studies where the reversal of rf current at a critical radial position was confirmed. The position of the critical radius agreed with predictions for the above ranges of parameters.

Conclusion

Evidence has been presented to show that Feenberg's small signal space charge theory is valid over a sufficiently wide range to be applicable to meet all reasonable tube designs. Moreover the extrapolation of this theory is valid for all values of bunching parameter used in conventional tubes. In all cases, it is found that the output radio frequency current decreases for increasing values of *hl*, and the voltage at the input required to maximize the output current increases with *hl* for all reasonable tube designs. It is important to note that voltages at both gaps were kept well below the beam voltage. The normalized input voltage was at most $\alpha=.3$ so that small signal theory can be safely applied to the input. The output voltage was kept small to simplify the evaluation of the rf current only. Since the theoretically predicted current at the output gap was obtained, the velocity and time of arrival of electrons at this gap are probably correctly predicted by the theory. The above knowledge makes it possible to obtain valid results for large signals at the output by merely integrating the equations of motion of the electrons through the gap to get the net transfer of power for any given rf voltage. It would be interesting to confirm this experimentally. It is indeed fortunate that this theory, and its extension, can be trusted under conditions for which it was not intended.

Acknowledgment

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A Simple Graphical Analysis of a Two-Port Waveguide Junction[*]

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Summary—**A graphical analysis based on the original work of Deschamps is presented for obtaining the scattering matrix of a two-port waveguide junction from standing-wave measurements. The section may have losses, and can be asymmetrical. In addition, a method is outlined whereby the reflection coefficient of a load terminating the junction can be obtained graphically from the measurement of the reflection coefficient as seen through the junction.**

INTRODUCTION

A GRAPHICAL METHOD was recently introduced by Deschamps[1] for obtaining the scattering matrix of a two-port waveguide junction from a set of standing-wave measurements. He also outlined a procedure whereby the reflection coefficient of a load terminating the junction can then be obtained graphically from a measurement of the reflection coefficient of the load as seen through the junction. (Some of the ideas involved have also been noted previously.[2]) The proofs given by Deschamps were largely in terms of projective and non-Euclidean geometry.

Deschamps' method requires less calculation than the direct algebraic computation of the scattering matrix from the measurements, and fewer measurements than using the Weissfloch-tangent technique[3] to obtain the equivalent circuit of the junction. The new method is general, and can be used without modification for lossy, asymmetrical junctions. Also, in its application, certain checks are set up for the elimination of errors of measurement.

Since this new method is of importance to engineers unfamiliar with higher geometry, its presentation using more familiar mathematics appears desirable. Certain additional features of the constructions also are discussed. The procedure for the use of the method is outlined first, with proofs and additional remarks given in the appendix.

I. PROCEDURE FOR OBTAINING THE SCATTERING MATRIX OF A TWO-PORT JUNCTION

Consider a junction between two guiding systems (transmission lines or waveguides) indicated symbolically in Fig. 1 as a two-port network. The two systems need not be identical, e.g., one side could feed into a two-wire line, the other into a coaxial line.

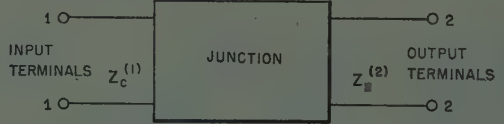

Fig. 1—Equivalent network representation of a two-port waveguide junction.

A. The Scattering Matrix

If the junction is equivalent to a linear passive network, its character can be described completely by the "scattering matrix" S of the junction, where

$$S = \left\| \begin{matrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{matrix} \right\|.$$

A typical matrix element S_{ij} has the usual interpretation of being the reflected ("scattered") wave at terminal i due to a unit-wave incident on the junction at terminal j.

Owing to reciprocity, $S_{12}=S_{21}$, and $S_{12}=T$, where T is the transmission coefficient of the junction. The impedance matrix is related to the scattering matrix by[4]

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[†] Cruft Laboratory, Harvard University, Cambridge, Mass.

[‡] RCA Fellow in Electronics under the National Research Council, Harvard University.

[1] G. Deschamps, "Application of Non-Euclidean Geometry to the Analysis of Waveguide Junctions," URSI-IRE Spring Meeting, 1952. Also in sections of Interim Reports 1–8 and Final Report on "Near-Zone Electromagnetic Waves," Federal Telecommunications Laboratories, Inc., Contract AF 33-038-13289.

[2] H. Wheeler and D. Dettinger, "Measuring the efficiency of a superheterodyne converter by the input impedance circle diagram," Wheeler Monographs, no. 9; March 1949. H. Wheeler, "The transmission efficiency of linear networks and frequency changers," Wheeler Monographs, no. 10; May 1949.

[3] N. Marcuvitz, "On the representation and measurement of waveguide discontinuities," PROC. I.R.E., vol. 36, pp. 728–735; 1948; also, A. A. Oliner and H. Kurss, "The Precision Measurements of the Equivalent Circuit Parameters of Dissipative Microwave Structures," Paper no. 78, I.R.E. National Convention, New York, N. Y.; 1951.

[4] C. G. Montgomery, R. H. Dicke, and E. M. Purcell, *Principles of Microwave Circuits*, vol. 8, pp. 146–149, Radiation Laboratory Series, McGraw-Hill, New York; 1948.

$$Z = \left\| \begin{matrix} \sqrt{Z_c^{(1)}} & 0 \\ 0 & \sqrt{Z_c^{(2)}} \end{matrix} \right\| \, \| I - S \|^{-1} \| I + S \| \, \left\| \begin{matrix} \sqrt{Z_c^{(1)}} & 0 \\ 0 & \sqrt{Z_c^{(2)}} \end{matrix} \right\|,$$

where $Z_c^{(1)}$ and $Z_c^{(2)}$ are the characteristic impedances of the two lines. Thus a knowledge of S is equivalent to a knowledge of Z. The scattering (or impedance) matrix of a two-port junction is usually obtained by terminating one side of the junction in various lengths of short-circuited line and measuring the reflection coefficients as seen from the other side.

In particular, by making such measurements for three different positions of the short circuit, it is possible to compute the scattering matrix of the junction by solving a system of three linear algebraic equations for the three unknown matrix elements. To insure accurate results (i.e., to guard against experimental errors), it is necessary to consider more positions of the short circuit, and essentially recalculate the values of these matrix elements, comparing them with those previously obtained.

Alternatively, by taking measurements for many positions of the short circuit, an equivalent circuit may be obtained by a modification of the Weissfloch-tangent technique[3] which, however, involves considerable labor.

Compared to these two procedures, Deschamps' approach is a simple and much less time-consuming graphical technique for obtaining the S-matrix elements from construction lines and angles. Measurements are needed for only four positions of the short circuit (6 or 8 may be used for increased accuracy), with checks possible for consistency of the data.

B. Graphical Evaluation of S_{11}

Suppose the values of the measured reflection coefficient $\Gamma' = \gamma' \epsilon^{+i\phi'}$ are plotted in the complex plane. (The harmonic time dependence $\epsilon^{+i\omega t}$ has been assumed.) Since by physical considerations $0 \leqq \gamma' \leqq 1$, all possible values lie within the unit circle, and the diagram is conveniently drawn on the usual Smith chart if the reflection-coefficient circles are available. (The resistance and reactance lines of the chart are of no concern whatever in this method.) Moreover, it is well known (see appendix) that when the output of the junction is connected to an arbitrary load, through a length l of lossless line, the values of reflection coefficient measured at the junction input lie on a circle, Fig. 2, when l is varied.

Deschamps' method proceeds as follows: The output port of the junction is connected to a short-circuited line of electrical length kl, and the complex reflection coefficients seen at the input port are measured for four lengths of line; these lengths are taken in pairs, kl_1, $kl_1+\pi/2$; kl_2, $kl_2+\pi/2$. (A good distribution of the data may be achieved conveniently by placing the short circuit at four positions equally spaced along a half-wavelength of the output line.)

These reflection coefficients are plotted in the complex plane (points P_n' in Fig. 2). The points $P_1'(kl=kl_1)$ and $P_3'(kl=kl_1+\pi/2)$ are connected by a straight line, as are $P_2'(kl=kl_2)$ and $P_4'(kl=kl_2+\pi/2)$; the intersection of these lines is labeled S_{11}' and termed the "cross-over point." Also a circle G' is fitted to the P_n' with center C' located, for example, as the intersection of the perpendicular bisectors of the chords $P_1'P_3'$ and $P_2'P_4'$.

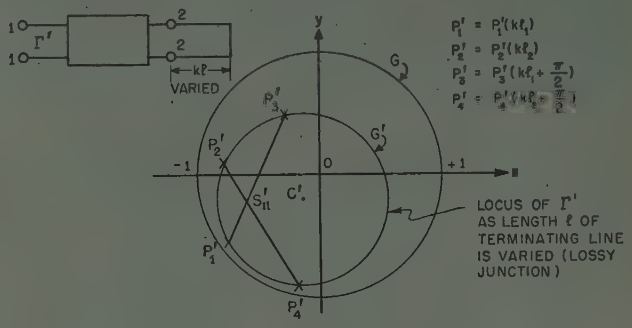

1
1
Γ'
2
2
kℓ
VARIED
y
G
G'
P'3
P'2
P'1
P'4
-1
0
+1
S'11
C'
$P_1' = P_1'(k\ell_1)$
$P_2' = P_2'(k\ell_2)$
$P_3' = P_3'(k\ell_1 + \frac{\pi}{2})$
$P_4' = P_4'(k\ell_2 + \frac{\pi}{2})$
LOCUS OF Γ' AS LENGTH ℓ OF TERMINATING LINE IS VARIED (LOSSY JUNCTION)

Fig. 2—Plot of measured reflection coefficients.

The line $S_{11}'C'$ is drawn, and perpendiculars $S_{11}'A'$ and $C'B'$ are erected as shown in Fig. 3. The intersection of $A'B'$ with $S_{11}'C'$ is labeled S_{11} and termed the "iconocenter." *The complex number corresponding to this point in the complex plane is the coefficient S_{11} in magnitude and phase.* Thus the reflection coefficient when the junction is matched at its output is already determined.

$$|S_{11}| = \overline{OS_{11}} \qquad \text{Arg } S_{11} = \measuredangle(\overrightarrow{OP}, \overrightarrow{OS_{11}})$$

$$|S_{12}| = \frac{\overline{S_{11}E'}}{\sqrt{r}} \qquad \text{Arg } S_{12} = \frac{1}{2} \measuredangle(\overrightarrow{OP}, \overrightarrow{C'P_3''})$$

$$|S_{22}| = \frac{\overline{S_{11}C'}}{r} \qquad \text{Arg } S_{22} = \measuredangle(S_{11}C', \overrightarrow{C'P_3''})$$

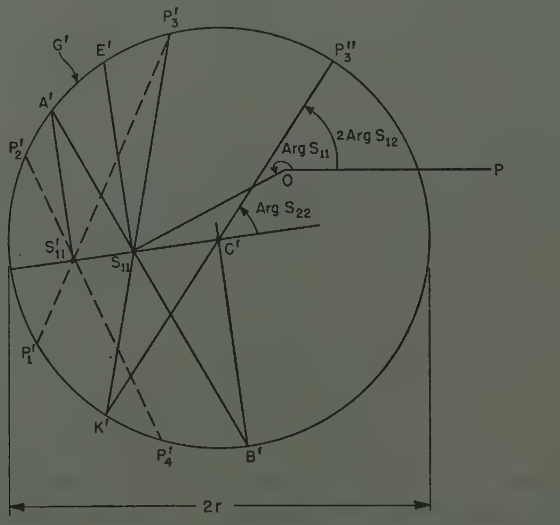


Fig. 3—Construction for elements of the scattering matrix.

C. Graphical Method—Evaluation of S_{12} and S_{22}

For the determination of the other elements of S, alternative procedures (i.e., the measurement of lengths and angles different from those used by Deschamps) can be devised easily by simple geometric considerations, but Deschamps' original procedure, as follows, thus far appears to be the simplest.

It is convenient to know the point on the measured circle which corresponds to an open circuit at the output reference plane (i.e., the short circuit placed $\lambda/4$ away). Assume first that this point happens to be one of the measured points, say P_3'. Then draw the line $P_3'K'$ through S_{11}, and the diameter $K'P_3''$. Also construct $S_{11}E' \perp S_{11}'C'$ shown in Fig. 3. Then, with r the radius of the measured circle, the values for the matrix elements are given by

	Absolute Value	*Phase Angle (directed)*
S_{11}	$\overline{OS_{11}}$	$\sphericalangle(\overrightarrow{OP, OS_{11}})$
S_{12}	$\overline{S_{11}E'}/\sqrt{r}$	$\frac{1}{2}\sphericalangle(\overrightarrow{OP, C'P_3''})$
S_{22}	$\overline{S_{11}C'}/r$	$\sphericalangle(\overrightarrow{S_{11}C', C'P_3''})$

The barred quantities are the lengths of the line segments. The angles are directed angles in the sense that the two directed-line segments which include the angle are given, and the angle is that through which the first segment is rotated into the second; the sign is positive if the rotation is counterclockwise, negative if clockwise. (Either direction of rotation may be used, the resulting angles then differing by 2π, which is of no electrical significance. There is also an indeterminacy by π in the phase of S_{12}, which is also of no electrical significance.)

If the point corresponding to an open circuit at terminal 2 is not one of the measured points, then any one of the measured points, say P_3' again, may be used in the construction. The magnitudes are still found as before, but the phase angles for S_{12} and S_{22} must be corrected as follows: If P_3' corresponds to a short circuit located at an electrical distance kl from the terminals of the junction, while a short circuit located at kl_0 gives the desired open circuit at these terminals, then the correction angle is $\alpha = 2(kl - kl_0)$. In terms of this angle,

	Phase angle (directed)
S_{12}	$\frac{1}{2}[\sphericalangle(\overrightarrow{OP, C'P_3''}) + \alpha]$
S_{22}	$[\sphericalangle(\overrightarrow{S_{11}C', C'P_3''}) + \alpha]$.

One of the most desirable features of obtaining the scattering matrix in this manner is that it is possible to obtain quickly an estimate of errors involved in the experimental data. This is discussed in Section III in conjunction with an illustrative example.

It is interesting that if the measurements are taken with an imperfect short circuit (or for that matter with any *arbitrary* load at the end of a lossless line of variable length, connected to the output port of the junction) and the constructions are carried out as described, the values obtained for S_{11} and the ratio S_{12}^2/S_{22} are still correct.[5]

II. Procedure for Measuring an Obstacle Through a Junction

Measurements made through a discontinuity, such as an adapting section, sometimes cannot be avoided. However, the section in which the adaptation occurs may be viewed as a two-port junction, and the unknown impedance can then be determined from measurement by simply adding certain constructions to those of Fig. 3. Two alternative procedures are available at the moment. However, the first, originally proposed by Deschamps,[1] is not discussed since the alternative that will be described seems simpler and more rapid.

The usual measurement is made: The unknown reflection coefficient Γ_x is connected at the output terminals of the junction for which the circle G' and points C', S_{11}', S_{11} have already been determined, as in Fig. 3; then the reflection coefficient Γ_x' at the input-reference plane is measured as some complex number, $\Gamma_x' = \gamma_x' \epsilon^{j\phi'_x}$, and plotted. (Although the work might well be superimposed on the previous diagram, a new diagram, Fig. 4, is drawn for clarity, and only the relevant points retained.)

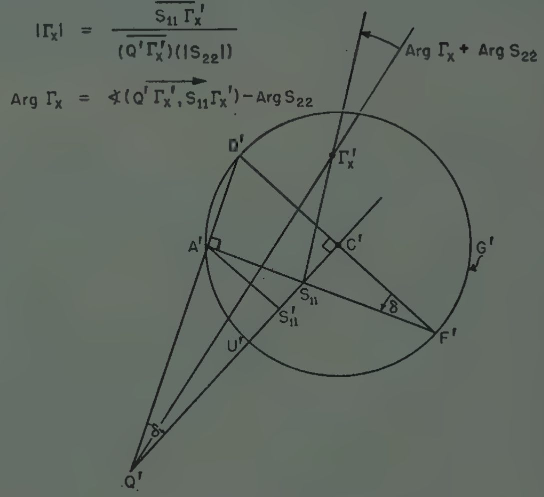

$|\Gamma_x| = \frac{\overline{S_{11}\Gamma_x'}}{(\overline{Q'\Gamma_x'})(|S_{22}|)}$
$\text{Arg}\,\Gamma_x = \sphericalangle(\overrightarrow{Q'\Gamma_x', S_{11}\Gamma_x'}) - \text{Arg}\,S_{22}$
Arg Γx + Arg S22
D'
Γx'
A'
C'
G'
S11
S11'
δ
U'
F'
δ
Q'

Fig. 4—Construction for Γ_x.

With this point determined, the basic procedure of the method, shown in Fig. 4, is as follows

Draw $S_{11}'A' \perp S_{11}'C'$, and $C'D' \perp S_{11}'C'$; draw $D'A'$ extended to meet $C'S_{11}$ (extended) at Q'; and draw $Q'\Gamma_x'$ and $S_{11}\Gamma_x'$. Then with $\Gamma_x = \gamma_x \epsilon^{+j\phi_x}$

$$|\Gamma_x = \gamma_x = \frac{\overline{S_{11}\Gamma_x'}}{(\overline{Q'\Gamma_x'})(|S_{22}|)}$$

$$\arg \Gamma_x = \phi_x = \phi - \arg S_{22} = \sphericalangle(\overrightarrow{Q'\Gamma_x', S_{11}\Gamma_x'}) - \arg S_{22}.$$

---

[5] See, for example, the authors' Technical Report no. 160, "A Simple Graphical Analysis of Waveguide Junctions," Cruft Laboratory, Harvard University; August, 1952.

Note that this method requires the knowledge of the junction parameter S_{22}.

Note also that as S_{11}' approaches close to C', the point Q' tends to fall off the sheet. In this case the following slightly different construction, which actually is useful in all cases, may be employed: Referring to Fig. 5, the lines $D'\Gamma_x'$ and $C'\Gamma_x'$ are drawn. From the point U' where $S_{11}'C'$ intersects the circle G', a line is drawn parallel to $A'D'$, intersecting $C'D'$ at V'. Through V' another line is drawn parallel to $D'\Gamma_x'$, and intersecting $C'\Gamma_x'$ in Y'. (E. W. Matthews, Cruft Laboratory, has pointed out that V' can be found more simply by marking off $C'V' = S_{11}C'$. The fact that $\overline{C'V} = \overline{S_{11}C'}$ can immediately be seen from proof, using Fig. 5, by noting that this is just

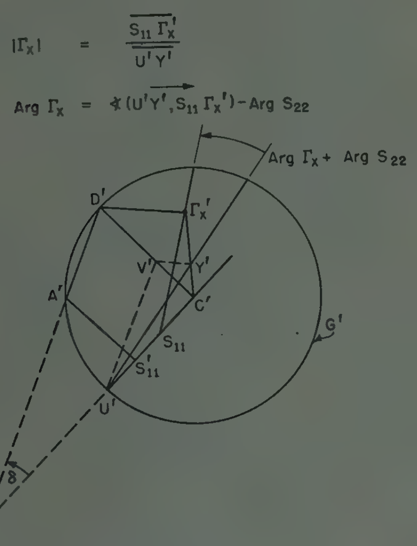

$|\Gamma_X| = \dfrac{\overline{S_{11}\Gamma_X'}}{\overline{U'Y'}}$
$\text{Arg}\ \Gamma_X = \sphericalangle(\overrightarrow{U'Y'}, S_{11}\Gamma_X') - \text{Arg}\ S_{22}$
Arg Γ_X + Arg S_{22}
D'
Γ_X'
V'
Y'
A'
C'
G'
S_{11}
S_{11}'
U'
δ
Q'

Fig. 5—Alternative construction for Γ_x.

$$\frac{\overline{C'V'}}{\overline{C'D'}} = \frac{\overline{C'V'}}{\overline{C'U'}} = \frac{\overline{S_{11}C'}}{r}.$$

The elimination of the one parallel-line construction greatly increases both the speed and the accuracy of the procedure.) Then (see the appendix, part F),

$$|\Gamma_x| = \gamma_x = \frac{\overline{S_{11}\Gamma_x'}}{\overline{U'Y'}}$$

$$\arg \Gamma_x = \phi_x = \sphericalangle\ (\overrightarrow{U'Y', S_{11}\Gamma_x'}) - \arg S_{22}.$$

III. APPLICATION TO AN ILLUSTRATIVE EXAMPLE

In most practical situations the power transmitted through, and reflected from, an arbitrary junction is of most interest. In addition, one is frequently faced with the problem of determining the impedance of an arbitrary load terminating a two-port junction located somewhere between the standing-wave detector and the load. In fact, in practically *all* direct measurements of load impedances there exist discontinuities (connectors, adapters, cables, and so on) between the standing wave detector and the load to be measured. The usual procedure involved in determining the impedance of a load through such a junction has been to determine first the impedance matrix characterizing the junction, and then to calculate the impedance of the load from the measured-input impedance to the combined junction and load. This procedure is tedious and requires the knowledge of *all* the impedance elements representing the four-terminal network of the junction. Using the simple graphical construction described above, the load impedance, usually the only quantity of interest, can be determined relatively quickly.

As illustration, a typical junction will now be analyzed and the elements of its scattering matrix determined graphically. An arbitrary load will then be placed at the output terminals of the junction and its impedance also determined graphically. Again, it is emphasized that the methods described are not unique and quite possibly simplifications may be made as the technique is developed in the future.

The experimental data were obtained from standing-wave measurements on a symmetrical lossy junction, although the symmetry was not necessary for use of the method. The characteristic impedance of the input and output transmission lines were equal. The impedance at the input terminals was measured with the short-circuiting plunger in the output line at positions of $kl = 0$, $\pi/8$, $\pi/4$, $3\pi/8$, $\pi/2$, $5\pi/8$, $3\pi/4$, $7\pi/8$, with $kl = 0$ corresponding to a short circuit at the output reference plane ($\Gamma = e^{-i\pi}$) and $kl = \pi/2$ an open circuit ($\Gamma = 1$).

Instead of only four data points, eight points were taken in order to provide additional accuracy against experimental errors. The positions of the short-circuiting pistons were spaced equally along a half-wavelength in the output line so as to obtain a good distribution of data.

The experimental values of impedance measured at the input terminals of the junction are plotted on a Smith chart in Fig. 6 (following page) for the indicated positions of the short-circuiting plunger in the output circuit. Since the constructions are to be carried out in the reflection-coefficient plane, it is advisable to use a Smith chart with superimposed reflection-coefficient circles. Obviously, the accuracy of the results depends on the accuracy of the constructions; hence, for precision measurements, an enlarged reflection-coefficient chart should be used. Here, for a desired three-place accuracy, graph paper one meter square was used.

Corresponding pairs of data points for $kl = (0, \pi/2)$, $(\pi/8, 5\pi/8)$, $(\pi/4, 3\pi/4)$ and $(3\pi/8, 7\pi/8)$ are joined by straight lines and the nature of intersection observed. Theoretically these chords should intersect in the crossover point S_{11}'. Owing to experimental error they may not, and the quality of their intersection is a measure of the consistency of the experimental data. Thus, if the four lines intersect in the manner shown in the enlarged

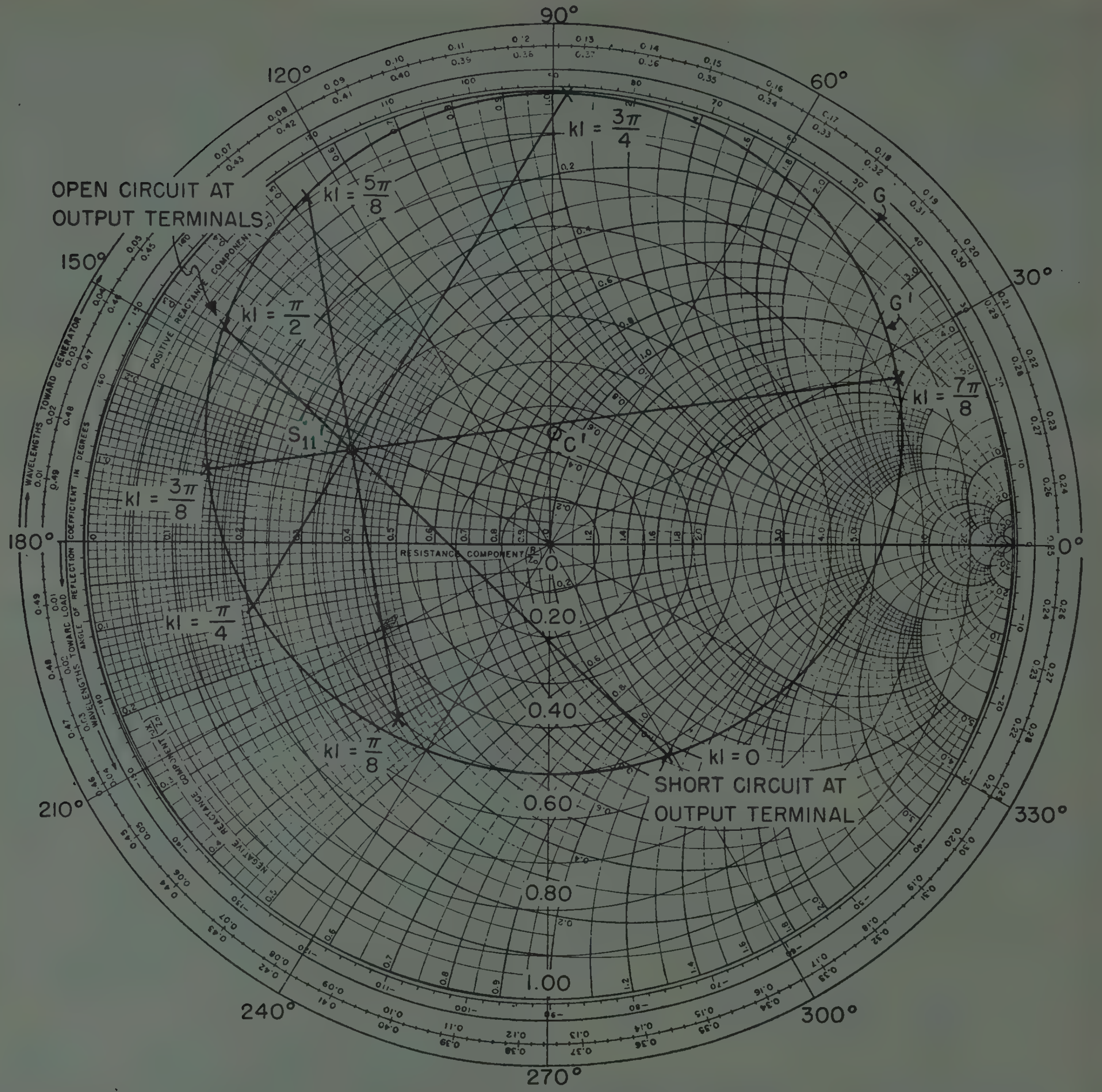


Fig. 6—Experimental data plotted on Smith-reflection coefficient chart and constructions for locating S_{11}' and C'.

view of the crossover region in Fig. 7, there is an indication of error in one of the two measurements associated with what appears to be a "bad" chord, i.e., for the shorting circuiting piston at either $kl=\pi/8$ or $5\pi/8$. Since two extra pairs of points have been taken, it appears reasonable to disregard this one line and take S_{11}' at the center of the tiny triangle formed by the other three lines. This analysis of the data can be performed directly as the readings are taken.

It has been shown that all measured points should lie on a circle of center C'. If the center of the circle is to be found by the intersection of the perpendicular bisectors of the various chords, a similar situation is encountered as in determining the crossover point. Again, only the six reliable points are used, and an appropriate center C' for the circle is chosen at the center of the triangle. Finally, the circle is drawn with an average radius, determined by averaging the distances from C' to the measured points. It is now apparent that the point corresponding to $kl=\pi/8$ is the one in error, probably owing to the fact that the standing-wave measurement was partially masked by noise.

Now that S_{11}' and C' have been found and the circle G' drawn, graphical constructions are performed as shown in Figs. 8 and 9 to determine the amplitudes and phases of the scattering matrix elements. Although these amplitude and phase constructions have been carried out here separately for clarity, in practice all can be per-

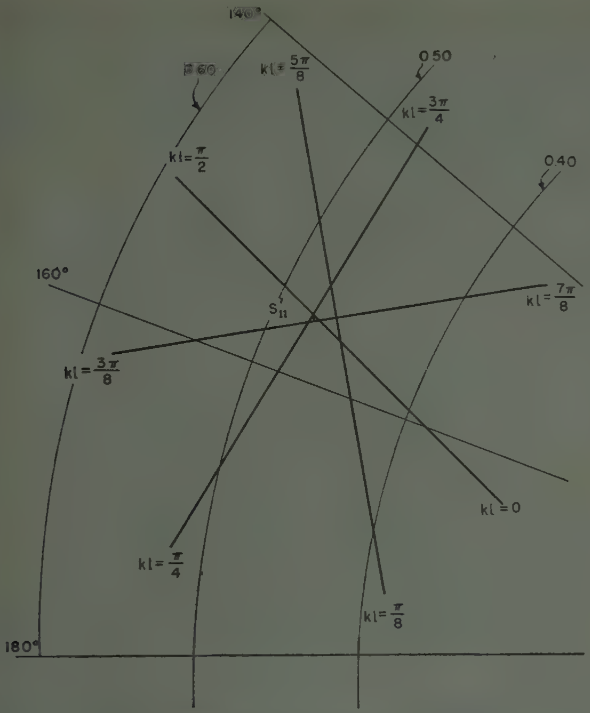

140°
0.50
150°
$kl=\frac{5\pi}{8}$
$kl=\frac{3\pi}{4}$
$kl=\frac{\pi}{2}$
0.40
160°
S_{11}
$kl=\frac{7\pi}{8}$
$kl=\frac{3\pi}{8}$
$kl=0$
$kl=\frac{\pi}{4}$
$kl=\frac{\pi}{8}$
180°

Fig. 7—Enlarged view of cross-over region.

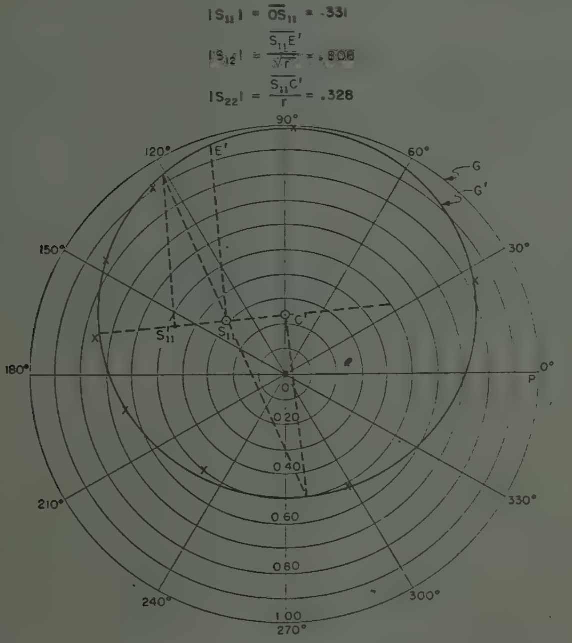

$|S_{11}| = \overline{OS_{11}} = .331$
$|S_{12}| = \frac{\overline{S_{11}E'}}{\sqrt{r}} = .808$
$|S_{22}| = \frac{\overline{S_{11}C'}}{r} = .328$
90°
120°
E'
60°
G
G'
150°
30°
S_{11}'
S_{11}
C'
180°
0°
P
0
0.20
0.40
210°
330°
0.60
0.80
240°
300°
1.00
270°

Fig. 8—Graphical determination of amplitudes of scattering matrix elements.

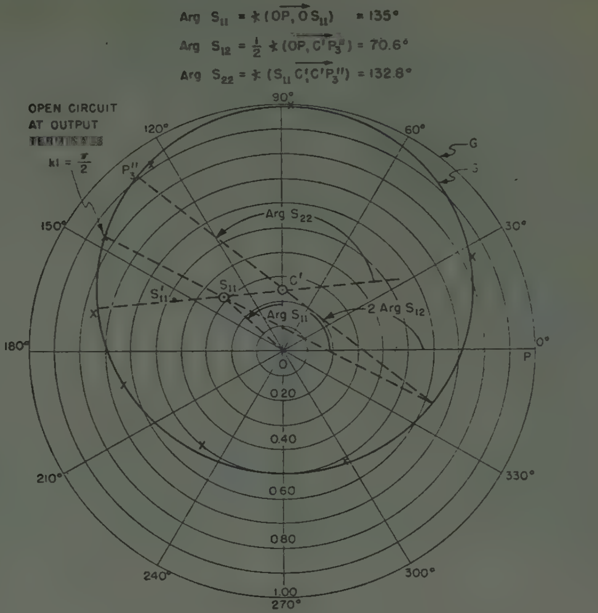

$\text{Arg } S_{11} = \measuredangle(\overrightarrow{OP}, \overrightarrow{OS_{11}}) = 135°$
$\text{Arg } S_{12} = \frac{1}{2} \measuredangle(\overrightarrow{OP}, \overrightarrow{C'P_3''}) = 70.6°$
$\text{Arg } S_{22} = \measuredangle(\overrightarrow{S_{11}C'}, \overrightarrow{C'P_3''}) = 132.8°$
90°
OPEN CIRCUIT
AT OUTPUT
TERMINALS
$kl=\frac{\pi}{2}$
120°
60°
G
G'
P_3''
150°
Arg S_{22}
30°
C'
S_{11}'
S_{11}
Arg S_{11}
2 Arg S_{12}
180°
0°
P
0
0.20
0.40
210°
330°
0.60
0.80
240°
300°
1.00
270°

Fig. 9—Graphical determination of phase angles of scattering matrix elements.

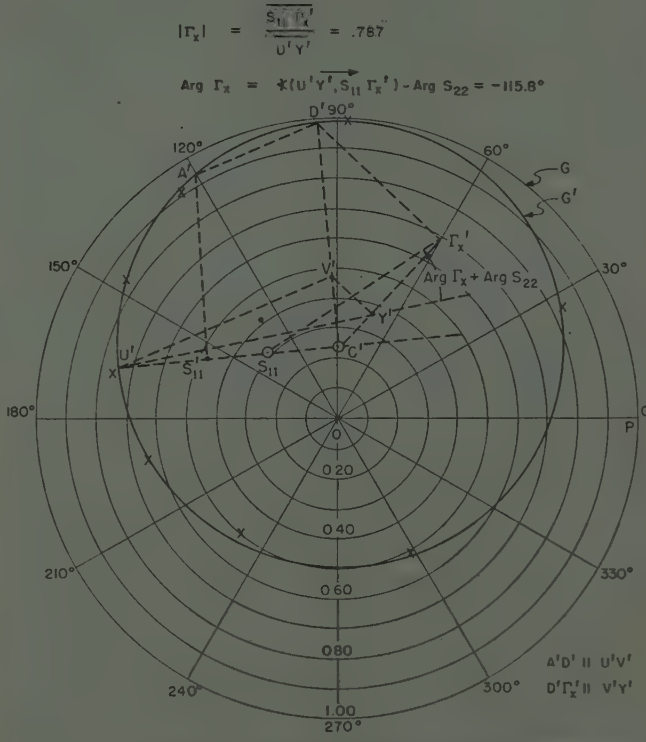

$|\Gamma_x| = \frac{\overline{S_{11}\Gamma_x'}}{\overline{U'Y'}} = .787$
$\text{Arg } \Gamma_x = \measuredangle(\overrightarrow{U'Y'}, \overrightarrow{S_{11}\Gamma_x'}) - \text{Arg } S_{22} = -115.8°$
D' 90°
120°
A'
60°
G
G'
Γ_x'
150°
V'
Arg Γ_x + Arg S_{22}
30°
Y'
U'
S_{11}'
S_{11}
C'
180°
0°
P
0
0.20
0.40
210°
330°
0.60
0.80
A'D' ‖ U'V'
D'Γ_x' ‖ V'Y'
240°
300°
1.00
270°

Fig. 10—Graphical method for obtaining reflection coefficient of a load terminating a symmetrical waveguide junction.

formed on the same sheet, and, with experience, many of the construction lines omitted.

Note that the amplitude results of Fig. 8 suffice to determine the division of power between the reflected wave and that transmitted, when the junction is terminated in a matched load. In this case, to calculate also a matching network at the input requires additionally only arg S_{11}, which is also available from Fig. 8 alone.

Next, an unknown impedance is used to terminate the junction, instead of the short-circuiting plunger. The reflection coefficient $\Gamma_x' = 0.70\epsilon^{j60°}$ is measured at the input terminals and plotted in Fig. 10. After performing the indicated constructions the reflection coefficient is found to be

$$|\Gamma_x| = \gamma_x = \frac{\overline{S_{11}\Gamma_x'}}{\overline{U'Y'}} = 0.787$$

$$\arg \Gamma_x = \phi_x = \measuredangle\,(\overrightarrow{U'Y', S_{11}\Gamma_x'}) - \underline{/S_{22}} = -115.8°.$$

The normalized load impedance can now be calculated as

$$Z_x = \frac{1+\Gamma_x}{1-\Gamma_x} = 0.165 - j0.615.$$

APPENDIX

A. *The junction as a bilinear transformation*

With the junction terminated in a load of reflection coefficient Γ, the reflection coefficient Γ' measured at the input port is easily expressed in terms of Γ, using the definitions of the elements of the scattering matrix, by

$$\Gamma' = S_{11} + \frac{S_{12}^2\Gamma}{1 - S_{22}\Gamma} = \frac{S_{11} + (S_{12}^2 - S_{11}S_{22})\Gamma}{1 - S_{22}\Gamma}. \quad (1)$$

This relation between the complex numbers Γ' and Γ is therefore of the form of a bilinear transformation, of which certain general properties are well-known.[6] For example: a) with straight lines considered as limiting cases of circles, the bilinear transformation (1) will map circles in the Γ-plane into other circles in the Γ'-plane, though generally with a different radii and centers; b) the transformation is conformal, i.e., if two curves intersect in Γ at a certain angle, their images will intersect in Γ' at the same angle, and with the same sense.

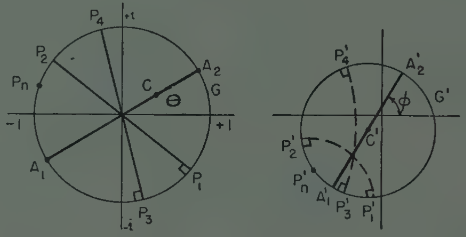
Now when the output to the junction consists of a load characterized by $\Gamma_L = \gamma_L \epsilon^{j\phi_L}$, viewed through a section of lossless line of electrical length kl, the reflection coefficient terminating the junction is

$$\Gamma = \gamma_L \epsilon^{j(\phi_L - 2kl)}.$$

The plot of this Γ in the complex plane as l is varied is a circle G of radius γ_L, centered at the origin. Hence when this circle is mapped by (1) into the Γ'-plane (i.e., when measurements are made through the junction), the image points will lie on some circle G' of center C'; and, since any diameter of G cuts G orthogonally, its image in w must be an arc passing through O' (image of the origin O) and intersecting G' orthogonally (Fig. 11). *Every diameter* is thus such an arc. The particular case of interest at the moment is when the load is a short circuit, $\Gamma_L = -1$, and G therefore the unit circle.

Now from (1) the image of $\Gamma = 0$ is $\Gamma' = S_{11}$. This point, labeled O', or more suggestively, S_{11}, therefore locates the complex number representing the matrix element S_{11} in both magnitude and phase. How to determine this "iconocenter" from the measured points located around G' will be discussed later; for the moment, let it be assumed to be located. Then, in Fig. 11, the diameter $A_1'C'A_2'$ through S_{11} is a particular case of an arc through O', orthogonal to G'; hence its original A_1A_2 must also have been a diameter of G. Putting $\Gamma = \rho\epsilon^{j\theta}$ along A_1A_2 and $S_{22} = |S_{22}|\epsilon^{j \arg S_{22}}$, the image of A_1A_2 is seen from (1) to be

$$\Gamma' = S_{11} + \frac{S_{12}^2\rho\epsilon^{j\theta}}{1 - \rho|S_{22}|\epsilon^{j(\theta + \arg S_{22})}} \quad \theta = \text{constant}, \rho \text{ varying}.$$



(a) (b)

Fig. 11—(a) Original circle in Γ′-plane (b) Image circle in Γ′-plane.

This in turn, however, can only be the straight line in Γ', as ρ varies, if A_1A_2 is so oriented that

$$\theta = -\arg S_{22}. \quad (3)$$

Hence points along the particular diameter $A_1'S_{11}A_2'$ are given parametrically by

$$\Gamma' = S_{11} + \frac{S_{12}^2\rho\epsilon^{j\theta}}{1 - \rho|S_{22}|}. \quad (4)$$

The end points A_1 and A_2 (lying on the unit circle) are characterized by $\rho = \mp 1$. Hence the radius r of G' can be determined as

$$r = \frac{1}{2}|\Gamma'_{A_2'} - \Gamma'_{A_1'}| = \frac{1}{2}\left|\frac{S_{12}^2}{1 - |S_{22}|} + \frac{S_{12}^2}{1 + |S_{22}|}\right| = \frac{|S_{12}|^2}{1 - |S_{22}|^2}. \quad (5)$$

Further, since $\Gamma'_{O'} = S_{11}$, the length $\overline{O'A_1'}$ (or $\overline{S_{11}A_1'}$) is

$$\overline{S_{11}A_1'} = |\Gamma'_{A_1'} - S_{11}| = \frac{|S_{12}|^2}{1 + |S_{22}|}. \quad (6)$$

Finally, noting that

$$\overline{S_{11}C'} = r - \overline{S_{11}A_1'}, \quad (7)$$

it follows that

$$\frac{\overline{S_{11}C'}}{r} = 1 - \frac{\overline{S_{11}A_1'}}{r} = |S_{22}|. \quad (8)$$

Thus, if the point S_{11} is known, the construction for $|S_{22}|$ is given by (8).

---

[6] R. V. Churchill, "Introduction to Complex Variables," McGraw-Hill, New York, N. Y.; 1948.

Further, using (8), (5) may be solved for

$$|S_{12}|^2 = r[1 - |S_{22}|^2]$$
$$= \frac{1}{r}[r^2 - r^2|S_{22}|^2] = \frac{1}{r}[r^2 - \overline{S_{11}C'^2}]$$
$$= \frac{1}{r}(\overline{S_{11}E'^2})$$

where $S_{11}E'$ is drawn as in Fig. 3.

The result that

$$|S_{12}| = \frac{\overline{S_{11}E'}}{\sqrt{r}} \tag{9}$$

follows immediately.

B. Some geometric theorems and constructions

Before examining the actual determination of S_{11}, or of the phase angles, it will be necessary to point out certain easily proved[5] geometric theorems and constructions.

Theorem 1: Let circles G_1 and G_2 intersect orthogonally at P, Q; and L be an arbitrary point on arc PQ of circle G_2, shown in Fig. 12. Draw QL and PL extended to M and N respectively. *Then line MN is a diameter of G_1, and is parallel to the tangent RS drawn at L to G_2.*

Construction 1: Given two circles G_1 and G_2 intersecting orthogonally at P, Q. A line may then be drawn through center C_1 of G_1, parallel to the tangent drawn to G_2 at any arbitrary point L of arc PQ. For, using Fig. 12, if QL is drawn extended to M, the diameter MC_1N is then the desired line, by theorem 1. (Note PLN is also a straight line.)

Construction 2: Given a diameter MN of circle G_1, and an arbitrary point L within the circle, using Fig. 12, then a circle G_2 may be drawn through L, such that G_1 and G_2 intersect orthogonally, and that the tangent to G_2 at L shall be parallel to MN. For, if ML is drawn extended to Q, and NL extended to P, the circle through P, L, Q is then the required G_2. (Three points uniquely determine a circle.)

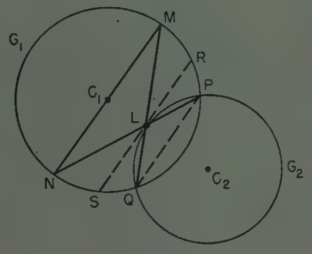


Fig. 12

Construction 3: If, in the above, LC_1 is perpendicular to MN, a slightly simpler construction is available, since now PQ is parallel to MN. Thus one can extend ML to Q, and obtain P by drawing PQ parallel to MN.

Theorem 2: Using Fig. 13, from an external point T, let TP be drawn tangent to a given circle G of center C (TP may also be regarded as the radius from center T, of a second circle intersecting G orthogonally at P).

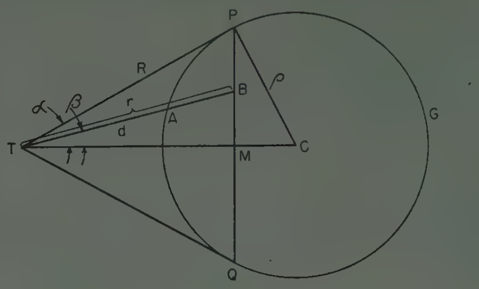


Fig. 13

Draw chord $PQ \perp TC$, cutting TC at M. Now draw *any* line from T, cutting arc PQ at A and chord PMQ at B. Then with $\overline{TA}=d$, $\overline{TB}=r$, $\overline{TP}=R$, the relation holds that

$$r = d\frac{2R^2}{R^2 + d^2}. \tag{10}$$

This relation between r, d, R holds in particular when A and B lie along the line TC. But then, given only a required value R for the length TP (but without prior knowledge of G), it will be possible to construct either A or B when the other is given (i.e., with R given, the relationship is *unique*). For, with reference to the interior of the circle centered at T, of radius $\overline{TP}$, the resemblance to the situation described in Construction 3 is obvious. This leads to

Construction 4: With Fig. 14, given a circle G_2 of center T, and an interior point A, by drawing diameter $MN \perp TA$, and applying Construction 3, the line PQ is obtained with the property that arc PAQ intersects G_2 orthogonally. But then radius TP is a tangent to the arc, and if TA (extended) intersects chord PQ at B, the relation between $\overline{TA}$, $\overline{TB}$, and $\overline{TP}$ is just that given in Theorem 4. Similarly, if B is given, A may be found by drawing diameter $MN \perp TB$, and chord $PBQ \perp TB$, and locating A as the intersection of MQ with TB.

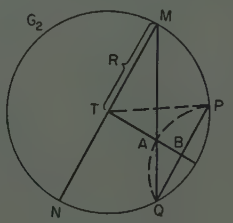


Fig. 14

The fact that in the above construction for locating (for a given circle) a point of character B from a point of character A, use has been made of (10) for the case when A and B lie on the line of centers between the two circles involved, *should not obscure* the fact that the ultimate result of Construction 4 is solely that TA, TB, and the radius R satisfy (10). That is, for any points A and B related to a circle by Construction 4, any chord drawn through B will determine an arc through A *necessarily* orthogonal to the circle. The relationship is unique, and Construction 4 is just its prescription.

C. The crossover property

As mentioned in part *A*, a short-circuited length of line has a reflection coefficient

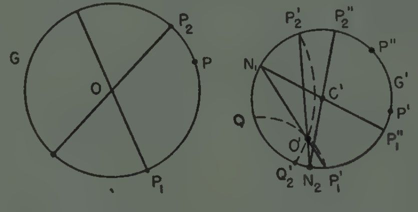
$$\Gamma = -\epsilon^{-j2kl},$$

and the values of Γ lie on the unit circle G as l is varied. Therefore, pairs of values of kl differing by $\pi/2$ represent points at opposite ends of diameters of G. Subsequently, when these values are transformed by the bilinear transformation (1), all diameters of G map into arcs passing through O' and intersecting G' orthogonally, as in Fig. 11. But then, by Construction 4, a point $\overline{O}$ can be found which lies on the chord of *any* such arc, and is hence a common point of intersection of all the chords. This is what has been termed the crossover point, and the general property may be stated as follows: Under any conformal transformation which maps circles into circles, the chords drawn in the image circle between the images of points originally at opposite ends of a diameter, all have a point in common; which point, further, is uniquely related to the image of the original center. (Note: This could have been shown more directly by noting that the diameter endpoints have the center as a center of negative inversion, and such a center must then also exist under the transformation; or, following Deschamps, by comparing stereographic and orthographic projections of a sphere onto its equatorial plane.)

The crossover property is the justification for the work in Section I, where first the crossover point, labeled there S_{11}', was located (knowing *a priori* of its existence) and then, by the method of Construction 4, the iconocenter S_{11} was located from S_{11}'.

D. "Reversion" of the image circle

Consider two points P_1' and P_2' on the image circle G', where P_1' and P_2' are images in the Γ'-plane of the points P_1 and P_2 in Γ of Fig. 15. If O' is the iconocenter, there are known to be arcs $P_1'O'Q_1'$ and $P_2'O'Q_2'$ orthogonal to G'. By Construction 1, applied as shown in Fig. 15 the diameters $N_1C'P_1''$ and $N_2C'P_2''$ are obtained, which, respectively, are parallel to the tangents of O' to these arcs. Therefore, $\sphericalangle P_1''C'P_2''$ included between these lines at C' is *equal* to the angle at which the arcs intersect at O'. But these arcs are themselves the images of the original diameters through P_1 and P_2, and by the angle-preserving properties of the bilinear transformation, intersect at the same angle as these original diameters. Therefore, $\sphericalangle P_1''C'P_2'' = \sphericalangle P_1OP_2$, so that the points P_1'' and P_2'' are located on the circumference of G' in the same relative positions as were the original points P_1 and P_2 on G. This process, of going from any P_n' to a P_n'' by use of Construction 1 and the particular point O' (the iconocenter) may be termed "reversion." If reversion is applied to each point of the measured (image) circle G', the resulting points (those double-primed in Fig. 15) will obviously then *all* be distributed around the circumference of G' in the same relative positions as were the corresponding original points (i.e., the points of G on which the bilinear transformation acted). Thus, the "reverted circle," the set of double-primed points may be thought of as a scaled-down, and rotated, version of the original circle G. Note that nothing has been said in this description about applying geometrical transformations to points of interior of G'.



(a) (b)

Fig. 15—(a) Original circle (b) image.

E. The elements of the scattering matrix

The phases of the S-matrix elements now may be calculated by using the reversion process. Referring to Fig. 11, it has been shown that the angle θ at which A_1A_2 is oriented in the original circle is equal to $-\arg S_{22}$. Hence, the latter may be determined by finding θ. Now, $\theta = \sphericalangle(\overrightarrow{OP, OA_2})$, where P is the point on the real axis in Γ at $\Gamma = +1$ (corresponding to an open circuit at the junction output). But A_2 maps into A_2' and P into a point P'. Suppose for the moment that the point P' has been one of the measured points. Then let reversion be applied to the points A_2' and P'. By its nature A_2' is unaffected, but P' goes into P'', as in Fig. 16, so that by the property of the reversion it follows that

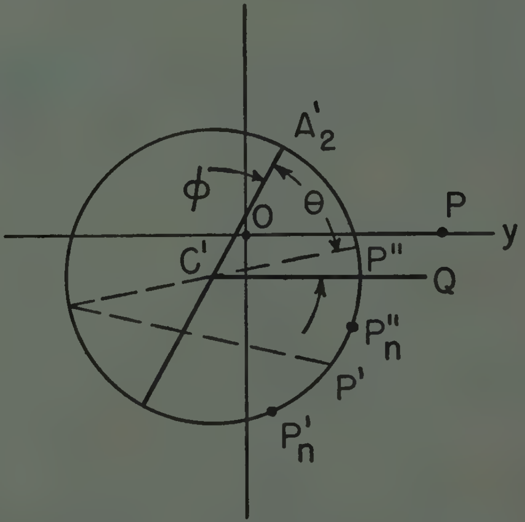


Fig. 16

$$\sphericalangle(\overrightarrow{C'P'', C'A_2'}) = \sphericalangle(\overrightarrow{OP, OA_2}) = \theta.$$

or

$$\arg S_{22} = -\theta = \sphericalangle(\overrightarrow{C'A_2', C'P''})$$
$$= \sphericalangle(\overrightarrow{S_{11}C', C'P''}).$$

The last step follows by inspection.

Next, it follows from (4) that the orientation of $A_1'A_2'$ with respect to the horizontal axis in Γ' is given by the angle ϕ in Fig. 16, where $\phi = 2 \arg S_{12} + \theta$. Or $\arg S_{12} = \frac{1}{2}(\phi - \theta)$, so that if on this diagram Q is a point on the horizontal line through C', $\arg S_{12} = \frac{1}{2}[\sphericalangle(\overrightarrow{C'Q,\ CP''})]$. If the real axis of this complex plane also has been labeled OP, so that OP and $C'Q$ are parallel, then

$$\arg S_{12} = \tfrac{1}{2} \sphericalangle (\overrightarrow{OP, C'P''}).$$

These results are given in Section I for the case when P' is one of the measured points (short circuit placed $\lambda/4$ past the junction).

If P' is unavailable and, instead, an arbitrary one of the measured points, P_n' is reverted into P_n'' and the angles measured, the difference is just that of the angle α included between $C'P''$ and $C'P_n''$. But this angle is that between the originals of these points in the Γ-plane. With P_n corresponding to the use of a length l of lossless line and P to a length l_0, it follows that

$$\alpha = \sphericalangle(\overrightarrow{C'P_n'', C'P''}) = \arg(\epsilon^{-j2kl_0}) - \arg(\epsilon^{-j2kl}),$$

or

$$\alpha = 2(kl - kl_0).$$

This is the quantity which must be added to the measured angles in each case in order to get the correct values for $\arg S_{22}$ and $\arg S_{12}$.

F. Measurements of an impedance through the junction

The procedure originally proposed by Deschamps makes extensive use of the reversion constructions. A simpler and quicker method has been outlined in Section II, and only this method will be discussed.

The proof proceeds from (1), applied to the case of a junction terminated in an unknown load Γ_x; Γ_x' is measured, giving the value

$$\Gamma_x' = S_{11} + \frac{S_{12}^2\Gamma_x}{1 - S_{22}\Gamma_x}\cdot \tag{11}$$

This equation may be solved for Γ_x, yielding

$$\Gamma_x = \frac{1}{S_{22}}\left[\frac{\Gamma_x' - S_{11}}{\Gamma_x' - \left(S_{11} - \dfrac{S_{12}^2}{S_{22}}\right)}\right]. \tag{12}$$

Suppose now that Γ_x' is plotted in the complex Γ'-plane, and also as point Q' the complex number $[S_{11} - (S_{12}^2/S_{22})]$, as shown in Fig. 17. Then, with

$$\Gamma_x = \gamma_x \epsilon^{j\phi}x$$

$$|\Gamma_x| = \gamma_x = \frac{1}{|S_{22}|}\,\frac{|\Gamma_x' - S_{11}|}{|\Gamma_x' - Q'|}$$

$$= \frac{\overline{S_{11}\Gamma_x'}}{|S_{22}|\cdot Q'\Gamma_x'},$$

and

$$\arg \Gamma_x = \phi_x = \arg(\Gamma_x' - S_{11}) - \arg(\Gamma_x' - Q') - \arg S_{22}$$
$$= \sphericalangle(\overrightarrow{Q'\Gamma_x', S_{11}\Gamma_x'}) - \arg S_{22}.$$

It remains to be shown that Q' is indeed the point obtained by the construction of Fig. 4. This is easily seen by noting that the point $\Gamma = \infty$ maps by (1) into just this value of $\Gamma' = S_{11} - (S_{12}^2/S_{22})$. But $\Gamma = 0$ and $\Gamma = \infty$ are inverse[6] with respect to the unit circle G in the Γ-plane and it is another property of bilinear transformations that their images (the points S_{11} and Q') must be inverse with respect to the image circle G'. That is, Q' can be located as the point on the radius through S_{11}, for which $(\overline{Q'C'})(\overline{S_{11}C'}) = r^2$, where r is the radius of G'. That the construction of Fig. 4 gives this result is readily shown from geometry.[5]

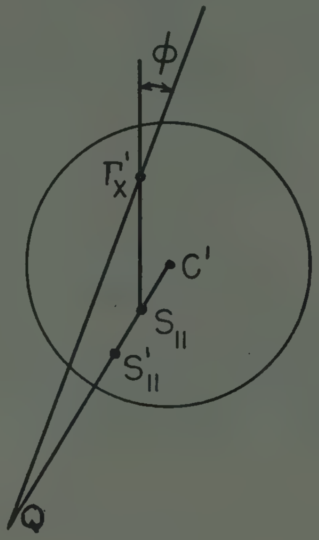


Fig. 17

As previously noted this method fails if Q' lies off the paper being used, i.e. if $S_{11}C'$ becomes too small.

However by proceeding from Fig. 4 to the construction in Fig. 5, a scaled-down version of the picture is obtained. Further, using the formula

$$|S_{22}| = \frac{\overline{S_{11}C'}}{r},$$

it may be noted from Figs. 4 and 5 that the scaling factor is just

$$\frac{\overline{C'U'}}{\overline{Q'C'}} = \frac{\overline{C'D'}}{\overline{Q'C'}} = \tan\delta = \frac{\overline{S_{11}C'}}{r} = |S_{22}|$$

and hence that

$$\frac{\overline{U'Y'}}{\overline{Q'\Gamma_x'}} = |S_{22}|,$$

or

$$\overline{U'Y'} = \overline{Q'\Gamma_x'}\cdot|S_{22}|; \tag{13}$$

and by the parallelism of $U'Y'$ and $Q'\Gamma_x'$,

$$\sphericalangle(\overrightarrow{U'Y', S_{11}\Gamma_x'}) = \sphericalangle(\overrightarrow{Q'\Gamma_x', S_{11}\Gamma_x'}). \tag{14}$$

The results claimed in Section II then follow immediately from (13), and (14).

The Gyrator as a 3-Terminal Element*

JACOB SHEKEL†, ASSOCIATE, IRE

Summary—**A 3-terminal circuit element, that has properties similar to those of a gyrator, is described. A method is derived to realize such an element with any unilateral transducer such as a vacuum-tube or a transistor. The effect of loading a gyrator by admittors in parallel or in series with its terminals is investigated. It is also shown that suitable loading of a gyrator may transform it into another gyrator of different gyrating admittance.**

INTRODUCTION

THE GYRATOR, a 2-terminal-pair network element that violates the theorem of reciprocity, has been postulated and analyzed by Tellegen.[1] Its properties are described by the "gyrating admittance" G, in

$$\left.\begin{aligned} I_1 &= GV_2 \\ I_2 &= -GV_1 \end{aligned}\right\}. \tag{1}$$

Given such an element, Tellegen proceeds to formulate a general method for synthesizing networks that may violate the reciprocity law.[2]

The physical realization of a gyrator seems to call for a medium that exhibits gyromagnetic properties. Such a medium has been used to construct a gyrator working in microwave frequencies.[3]

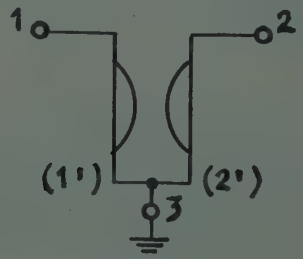


Fig. 1—Gyrator as a 3-terminal element.

If we short-circuit terminals 1′ and 2′ of the gyrator shown in Fig. 1, a 3-terminal element is obtained, whose admittance matrix, according to (1), is

$$\begin{Vmatrix} O & G \\ -G & O \end{Vmatrix}. \tag{2}$$

The 3-terminal element may be used with node 1 or 2 grounded. The indefinite-admittance matrix,[4] that is obtained from (2) by completing each row and column to zero, is

$$\begin{Vmatrix} O & G & -G \\ -G & O & G \\ G & -G & O \end{Vmatrix}. \tag{3}$$

When any terminal is grounded, the corresponding row and column are crossed out. It is obvious that, no matter which terminal is grounded, the resultant 2×2 matrix is always the same as (2) (with a possible reversal of sign). The 3-terminal gyrator exhibits the same properties in all grounding positions.

In view of this circular symmetry, the symbol of Fig. 2 is proposed to represent a 3-terminal gyrator. When the gyrator is incorporated in a network, elements $Y_{pq} = \pm G$ appear in the admittance matrix (p and q are two out of the three numbers i, j, k); $Y_{pq} = +G$ if pq is in the order indicated by the arrow, and $-G$ if pq is in the reverse order. This may also be described by a 3-index admittance symbol, indicating the order of rotation. Fig. 2 shows a gyrator of gyration admittance

$$Y_{ijk} = G, \tag{4}$$

meaning that, for instance, $Y_{ki} = G$ and $Y_{ji} = -G$.

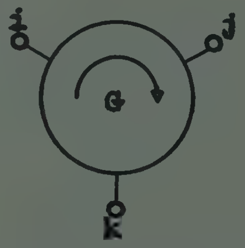


Fig. 2—Proposed symbol for 3-terminal gyrator.

PARALLEL-LOADED GYRATOR

Consider any linear 3-element network, active or passive, described by a 2×2 admittance matrix

$$\begin{Vmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{Vmatrix}, \tag{5}$$

with $Y_{12} \neq Y_{21}$. The matrix may be split into symmetric and skew-symmetric components

$$\begin{Vmatrix} Y_{11} & \frac{1}{2}(Y_{12}+Y_{21}) \\ \frac{1}{2}(Y_{12}+Y_{21}) & Y_{22} \end{Vmatrix} + \begin{Vmatrix} O & \frac{1}{2}(Y_{12}-Y_{21}) \\ \frac{1}{2}(Y_{21}-Y_{12}) & O \end{Vmatrix}. \tag{6}$$

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* Decimal classification: R143. Original manuscript received from the Institute, August 28, 1952; revised manuscript received, March 2, 1953.

† Scientific Department, Ministry of Defense of Israel; Haifa, Israel.

[1] B. D. H. Tellegen, "The gyrator, a new electric network element," *Phillips Res. Rep.*, 3, pp. 81–101; 1948.

[2] B. D. H. Tellegen, "The synthesis of passive, resistanceless four-poles that may violate the reciprocity relation," *Phillips Res. Rep.*, 3, pp. 321–337; 1948.

[3] C. L. Hogan, "The ferromagnetic Faraday effect at microwave frequencies and its applications—the microwave gyrator," *Bell Sys. Tech. Jour.*, vol. 31, p. 1; January, 1952.

[4] J. Shekel, "Matrix representation of transistor circuits," PROC. I.R.E., vol. 40, p. 1493; November, 1952.

The skew-symmetric part is a gyrator of gyration admittance

$$G = \tfrac{1}{2}(Y_{12} - Y_{21}), \tag{7}$$

and the symmetric part is a network composed of a triangle of admittors

$$\begin{aligned} A &= Y_{22} + \tfrac{1}{2}(Y_{12} + Y_{21}) \\ B &= Y_{11} + \tfrac{1}{2}(Y_{12} + Y_{21}) \\ C &= \qquad - \tfrac{1}{2}(Y_{12} + Y_{21}) \end{aligned} \tag{8}$$

the whole network being a parallel combination of both, as Fig. 3 shows. (In this discussion and throughout the rest of the paper, A, B, C, P and Q will represent admittances, in order not to encumber the equations with indexed Y's.)

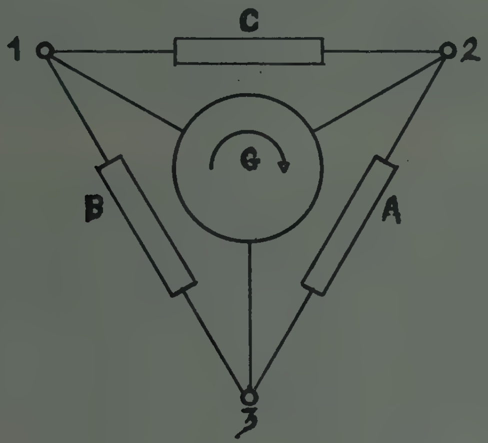


Fig. 3—The general 3-terminal network as a parallel-loaded gyrator.

It is now easy to perceive how an ideal, unloaded gyrator may be realized. It is only necessary to connect an admittance $-A$ in parallel with A, and in the same manner to "strip off" B and C. As negative resistors or other negative circuit elements are already acceptable circuit elements,[5] the realization of an ideal 3-terminal gyrator does not necessitate the postulation of any new components.

We may start with a triode operating in class A. Let G_m and G_p be its mutual conductance and plate conductance, respectively, and let the grid, anode and cathode be numbered in the order mentioned. The admittance matrix for grounded-cathode operation, split into symmetric and skew-symmetric components, is

$$\begin{Vmatrix} 0 & 0 \\ G_m & G_p \end{Vmatrix} = \begin{Vmatrix} 0 & \tfrac{1}{2}G_m \\ \tfrac{1}{2}G_m & G_p \end{Vmatrix} + \begin{Vmatrix} 0 & -\tfrac{1}{2}G_m \\ +\tfrac{1}{2}G_m & 0 \end{Vmatrix}, \tag{9}$$

showing that the triode may be regarded as a loaded gyrator, of gyration admittance

$$Y_{321} = \tfrac{1}{2}G_m. \tag{10}$$

The external loading necessary to "strip" the gyrator is shown in Fig. 4. It corresponds to a network whose admittance matrix is the negative of the symmetric matrix in (9).

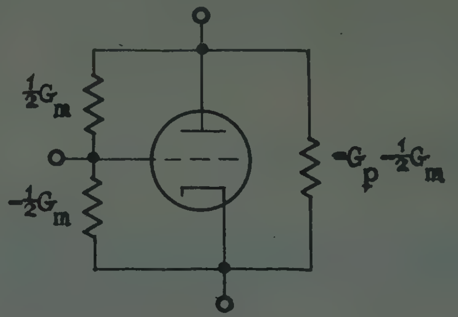
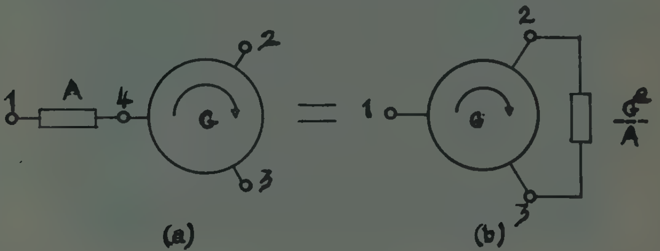


Fig. 4.—Realization of an ideal gyrator.

SERIES-LOADED GYRATOR

We will now investigate the properties of circuits that are composed of a gyrator, with admittors connected in series with some of its terminals.



(a) (b)

Fig. 5—Gyrator series-loaded with one admittor.

Fig. 5(a) shows a gyrator with one admittor A. The (indefinite) admittance matrix for this network is

$$\left\| \begin{array}{ccc:c} A & 0 & 0 & -A \\ 0 & 0 & G & -G \\ 0 & -G & 0 & G \\ \hdashline -A & G & -G & A \end{array} \right\|. \tag{11}$$

The matrix is written in a partitioned form, for we now want to eliminate the fourth terminal, and treat it as a 3-terminal network, with terminals 1, 2 and 3 only. The elimination[6] results in a 3×3 admittance matrix

$$\begin{Vmatrix} A & 0 & 0 \\ 0 & 0 & G \\ 0 & -G & 0 \end{Vmatrix} - \frac{1}{A} \times \begin{Vmatrix} -A \\ -G \\ G \end{Vmatrix} \times \begin{Vmatrix} -A & G & -G \end{Vmatrix}$$

$$= \begin{Vmatrix} 0 & G & -G \\ -G & \dfrac{G^2}{A} & G - \dfrac{G^2}{A} \\ G & -G - \dfrac{G^2}{A} & \dfrac{G^2}{A} \end{Vmatrix}. \tag{12}$$

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[5] H. W. Bode, "Network Analysis and Feedback Amplifier Design," pp. 185–188; Van Nostrand Publishing Co., New York, N. Y., 1945.

[6] G. Kron, "Tensor Analysis of Networks," Chap. X, "Reduction formulas," equations 10.2 and 10.7, pp. 243–244, Wiley and Sons Publishers, New York, N. Y.; 1939.

This is the same gyrator, parallel-loaded by an admittor G^2/A between terminals 2 and 3 as Fig. 5(b) shows. (This series-parallel transformation is also a property of the two-terminal-pair gyrator.)[1]

Consider now a gyrator with two admittors, A and B, in series with two of its terminals as shown in Fig. 6(a).

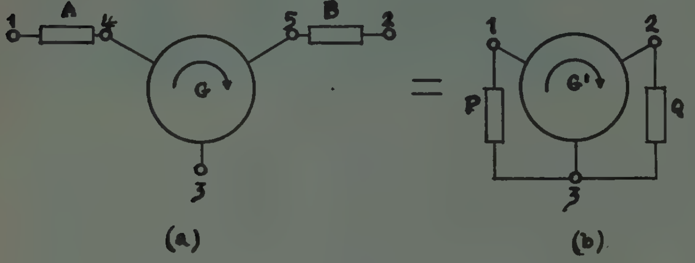

Fig. 6—Gyrator series-loaded with two admittors.

We will eliminate nodes 4 and 5, and thus find the equivalent network between nodes 1, 2 and 3. The complete indefinite admittance is

$$\left\|\begin{array}{ccc:cc} A & 0 & 0 & -A & 0 \\ 0 & B & 0 & 0 & -B \\ 0 & 0 & 0 & G & -G \\ \hdashline -A & 0 & -G & A & G \\ 0 & -B & G & -G & B \end{array}\right\|. \tag{13}$$

Elimination of nodes 4 and 5 results in the admittance matrix

$$\begin{aligned} &\left\|\begin{matrix} A & 0 & 0 \\ 0 & B & 0 \\ 0 & 0 & 0 \end{matrix}\right\| - \left\|\begin{matrix} -A & 0 \\ 0 & -B \\ G & -G \end{matrix}\right\| \times \left\|\begin{matrix} A & G \\ -G & B \end{matrix}\right\|^{-1} \times \left\|\begin{matrix} -A & 0 & -G \\ 0 & -B & G \end{matrix}\right\| \\ &= \left\|\begin{matrix} A & 0 & 0 \\ 0 & B & 0 \\ 0 & 0 & 0 \end{matrix}\right\| - \frac{1}{AB+G^2} \left\|\begin{matrix} -A & 0 \\ 0 & -B \\ G & -G \end{matrix}\right\| \times \left\|\begin{matrix} B & -G \\ G & A \end{matrix}\right\| \times \left\|\begin{matrix} -A & 0 & -G \\ 0 & -B & G \end{matrix}\right\| \\ &= \left\|\begin{matrix} A & 0 & 0 \\ 0 & B & 0 \\ 0 & 0 & 0 \end{matrix}\right\| - \frac{1}{AB+G^2} \times \left\|\begin{matrix} A^2B & -ABG & ABG+AG^2 \\ ABG & AB^2 & -ABG+BG^2 \\ -ABG+AG^2 & ABG+BG^2 & -AG^2-BG^2 \end{matrix}\right\| \\ &= \frac{G}{AB+G^2} \times \left\|\begin{matrix} AG & AB & -AB-AG \\ -AB & BG & AB-BG \\ AB-AG & -AB-BG & AG+BG \end{matrix}\right\|. \end{aligned} \tag{14}$$

Writing, for short,

$$\frac{ABG}{AB+G^2} = G'; \quad \frac{AG^2}{AB+G^2} = P; \quad \frac{BG^2}{AB+G^2} = Q \tag{15}$$

(14) reduces to

$$\left\|\begin{matrix} P & G' & -G'-P \\ -G' & Q & G'-Q \\ -P+G' & -Q-G' & P+Q \end{matrix}\right\|, \tag{16}$$

which is a gyrator G' parallel-loaded by P and Q as seen in Fig. 6(b).

The result may be applied to realize a gyrator of any gyrating admittance out of any given gyrator. If, in Fig. 6(a), an admittor $-P$ is connected between 1 and 3, and $-Q$ between 2 and 3, the result is an ideal, unloaded gyrator of admittance G'. (Given G and G', A and B are not determined uniquely, as only the product AB appears in the expression for G'.)

Series-loading all three terminals will give a similar result: a gyrator of different gyrating admittance, parallel-loaded by three admittors. The complete derivation of this result is not presented here, as it gives no new effects.

CONCLUSION

It was shown that a three-terminal element may be conceived, that has properties similar to those of the 2-terminal-pair gyrator. This element may be considered as the nucleus of any 3-terminal network that violates the reciprocity relation. Suitable loading of any such network, (e.g. vacuum-tube or transistor), by positive or negative bilateral admittors, may result in stripping the gyrator to its ideal, unloaded form.

Given a gyrator of any gyrating admittance, it is possible to realize gyrators of other admittance values by suitable series and parallel loading.

The necessary components for realizing a 3-terminal gyrator of any gyrating admittance are: a 3-terminal element or network that violates the reciprocity relation, the usual passive and bilateral network elements, and negative resistances.

A Method for Improving the Read-Around-Ratio in Cathode-Ray Storage Tubes*

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Summary—**Cathode-ray storage tubes are subject to interference which is caused by secondary electron flow between storage locations. A method of reducing this interference by modulating the cathode-ray tube differently during "action" and "regeneration" periods is described. This modulation builds up during "regeneration" periods a negative charge barrier around each spot. The negative charge barrier in turn reduces the flow of secondary electrons from one spot to adjacent spots during "action" periods. This results in decreased secondary electron flow from one storage location to another during "action" periods at the expense of increased secondary electron flow between storage locations during "regeneration" periods. Thus advantage is taken of the fact that practical computer stores can tolerate during "regeneration" periods redistribution currents between adjacent storage locations of the order of 100 times those flowing during "action" periods.**

I. INTRODUCTION

SINCE THE WORK of Williams and Kilburn[1] was published the ordinary cathode-ray tube has been widely adopted as a high-speed store for digital electronic computers. However, a serious defect of the cathode-ray tube store has become apparent, especially in parallel machines. When the area adjacent to a given storage location is consulted too frequently, the content of this storage location is obliterated. The maximum number of times spots adjacent to a given spot S may be consulted between regenerations of S without changing the information stored at S may be defined as the Read-Around-Ratio (R.A.R.) of S. In many machines the lowest R.A.R. has been found to be of the order of 20, while for foolproof performance a

$$\text{R.A.R.} \geqq \frac{\text{Number of spots in array}}{\text{Regenerations per action}}$$

is required. A R.A.R. between 100 and 1,000 is dictated by the above requirement for the optimum utilization of cathode-ray storage tubes in practical machines.

A method for increasing the R.A.R. is outlined below.

II. POTENTIALS OF THE FACE OF A CATHODE-RAY STORAGE TUBE

In a cathode-ray storage tube electrostatic charges are stored in a discrete array (raster) of spots, termed storage locations, shown in Fig. 1. By charging a storage location either directly with the primary electron beam, or indirectly with a secondary electron beam emanating from an area closely adjacent to the storage location proper, the storage location may be charged relatively positively or negatively. Three areas of the inside surface of the cathode-ray tube screen are of interest. These are

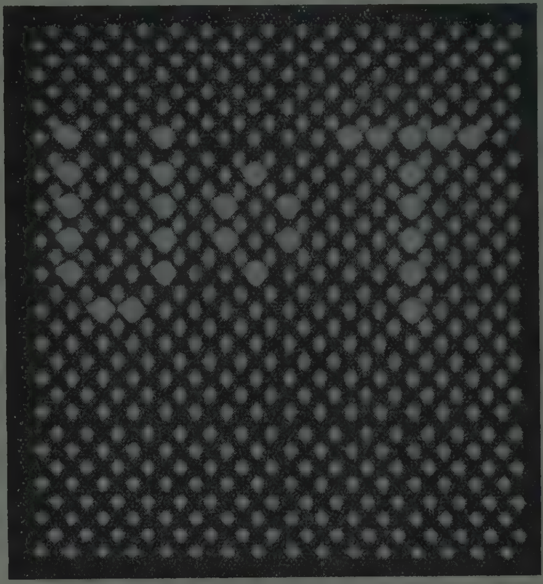

Fig. 1—Array of storage locations.

a) The storage locations proper.
b) The areas adjacent to the storage locations which are bombarded by the primary beam in order to furnish secondary electrons to the storage locations proper whenever the storage location is to be charged negatively. These areas will be termed secondary cathodes.
c) The remainder of the tube screen which serves neither for storage locations nor as secondary cathodes. This comprises usually the greater part of the screen area and will be termed the separation-area.

The potential diagram[2–4] of a cross-section of a cathode-ray storage-tube screen in the absence of beam current is indicated in Fig. 2. The space between lines a_1 and b_1 is a section of the storage location S_1. The space between b_1 and c_1 is a section of the secondary cathode of S_1. The space between c_1 and a_2 is a section of the

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[1] F. C. Williams and T. Kilburn, "A storage system for use with binary digital computing machines," *Proc. I.E.E.*, (London), part II, vol. 96; April 1949.

[2] L. Brillouin, "Theory of the Williams' Tube." Report on the 11th Annual Conference on Physical Electronics held March 29–31, 1951, at M.I.T.

[3] W. B. Nottingham, "Bombardment of Insulating Surfaces by Electron Beams." Report on the 11th Annual Conference on Physical Electronics, March 29–31, 1951, at M.I.T.

[4] J. Kates, "Space Charge Effects in Cathode-Ray Storage Tubes." Ph.D. Thesis, University of Toronto; 1951.

separation area between S_1 and S_2. The potential is referred to that of the electrode collecting the secondary electrons, usually the last anode of the tube. It will be noted that the separation area is at a negative potential with respect to the secondary electron collector. The storage locations and secondary cathodes are positive with respect to the separation area.

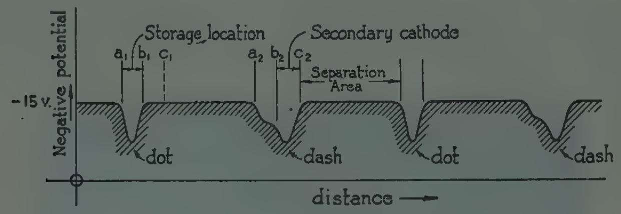

Fig. 2—Potential of a cross-section of the screen.

When the cathode-ray beam bombards a spot on the tube face, secondary electrons will leave this spot. The potential of the spot is governed by the potential drop in the secondary beam flowing to the secondary electron collector and, in a given tube, depends chiefly on the primary beam intensity. Some of the secondaries from the spot will flow to the separation area and to adjacent storage locations. The separation area will charge to a potential with respect to that of the storage locations and secondary cathodes, such that the integrated secondary electron current flowing to any spot of the separation area is equal to the integrated electron current leaking off the same spot. Since the secondary electrons leave the storage locations and secondary cathodes with a range of initial energies, the separation area has to charge negatively with respect to the area bombarded by the primary beam in order to limit the stray secondary electron current flowing to the separation area.

The full line of Fig. 3 is a potential diagram of a cross-section of a cathode-ray storage tube face while the primary beam bombards one spot. Primary and secondary electron paths have also been indicated. Vertical potential scale refers solely to the heavy potential lines. A vertical distance scale would refer to electron paths.

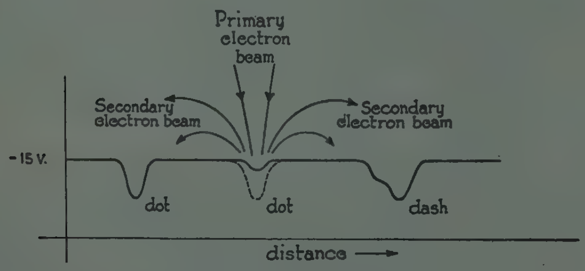

Fig. 3—Screen potential during bombardment.

The negative potential of the storage location and its immediate surroundings is in part due to the space charge in the electron beams traveling to and from the storage location. Hence, when the primary beam is removed, the potential of the storage location and its immediate surroundings will become more positive as indicated by the dotted line of Fig. 3.

III. “Action” and “Regeneration” Periods

The operation of a cathode-ray tube store is divided, not necessarily equally, into “action” and “regeneration” periods.[1] During action periods, the store furnishes some particular portion of its contents to other machine organs or places new contents into some particular place in the store.

Regeneration periods are interspersed with action periods to prevent the loss of information held in the store due to secondary electron redistribution, electrical leakage, stray ionization currents and so on. During successive regeneration periods, successive storage locations are scanned and their contents are restored.

Interference between storage locations is caused by the flow of secondary electrons from one storage location to another. This is aggravated by the fact, that during action periods the electron beam may bombard the immediate vicinity of a given storage location S a considerable number of times, between the times at which the given storage location S is regeneratd.

Interference could be greatly decreased if the secondary electron flow between adjacent storage locations could be decreased or prevented during action periods, even at expense of increased secondary electron flow between storage locations during regeneration periods. This result is achieved by method here proposed.

IV. Theory of the Method

The separation area surrounding a given storage location and its corresponding secondary cathode, will charge slightly negatively with respect to the most negative potential of the storage location and (or) the secondary cathode. The potential distribution will be such that the small-charge leakage from the separation area will just be replaced by secondary electrons from the area bombarded by primary electrons.

The principle of the method here proposed is to adjust conditions differently during action and regeneration periods, so that secondary electrons cannot flow to and over the separation area from a storage location or secondary cathode during action periods. This can be achieved by adding a so-called “guard” period which will usually be merged with the regeneration period. During the guard period the storage location and/or the secondary cathode or some portion of the separation area is driven to a more negative value than during any other period by

a) Intensifying the primary beam (i.e. increasing the beam current, adjusting the beam focus, or a combination of these methods). This increases the potential drop between the electrode or electrodes collecting or governing the secondary beam and the spot bombarded by the primary beam.
b) Pulsing negatively the electrodes or electrode collecting secondary electrons or governing flow.
c) Pulsing positively the pick-up electrode; or
d) A combination of some or all of a), b), and c).

The guard period will usually, but not necessarily, be merged with the regeneration period. During the guard period the storage location or the secondary cathodes or some portion of the separation area may be bombarded by the primary beam. The areas bombarded by the primary beam must be chosen so that the separation area is charged to an extent which diminishes or prevents secondary electron current flow between adjacent storage locations during action periods. Merging the guard period with the regeneration period achieves this automatically.

This method ensures that during the guard periods the separation area or a portion thereof is charged so negatively as to greatly diminish or prevent the flow of secondaries from one storage location or its secondary cathode to another storage location during action periods. The negative charge leaking off the separation area is replaced chiefly or only during the guard periods. This is indicated in Fig. 4 where the potential of a spot is shown during guard (full line) and action (dotted line) periods.

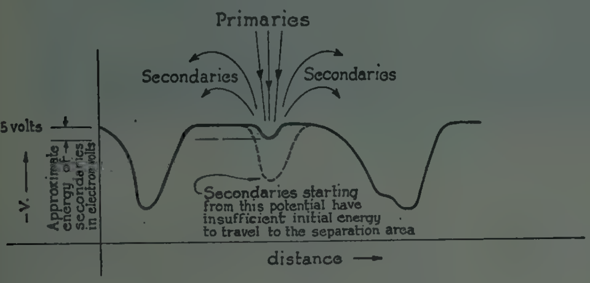
Primaries
Secondaries
Secondaries
-25 volts
-V.
Approximate energy of secondaries in electron volts
Secondaries starting from this potential have insufficient initial energy to travel to the separation area
distance

Fig. 4—Potential distribution during "guard" (full line) and "action" (dotted line) periods.

V. Examples of the Method

This section will illustrate a number of ways in which the method may be applied.

Fig. 5 shows the grid and deflection pulses of a cathode-ray storage tube during action, regeneration, and guard periods for both dots and dashes. If this system were to be applied to a defocus-focus system, the focus waveform would take the form of the deflection waveform shown in Fig. 5.

Another way of applying the method is shown in Fig. 6 where the potential of the post accelerator (third anode) is made negative during the guard period.

This may be combined with the beam modulation of Fig. 5. During the guard period the pick-up plate may be pulsed positively as shown in Fig. 7. This will have substantially the same result as the method of Fig. 6.

A way of effecting a similar result is to work with a defocused beam at all times except during the guard period at which time the beam is sharply focused. This is indicated in Fig. 8.

A combination of these methods may be used also. In every case the potential of a number of spots is depressed either directly (Figs. 5, 6, and 8) or indirectly, as Fig. 7 shows, while these spots are bombarded by the primary electron beam. The secondary electrons emanating from these spots charge the separation area so negatively that subsequently only the fastest secondaries from action spots can overcome the potential barrier so created.

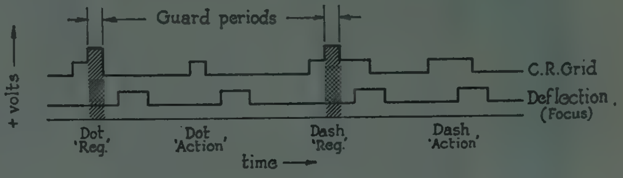
Guard periods
C.R.Grid
Deflection
(Focus)
+ volts
Dot 'Reg.'
Dot 'Action'
Dash 'Reg.'
Dash 'Action'
time

Fig. 5—Intensification of the beam during the guard period.

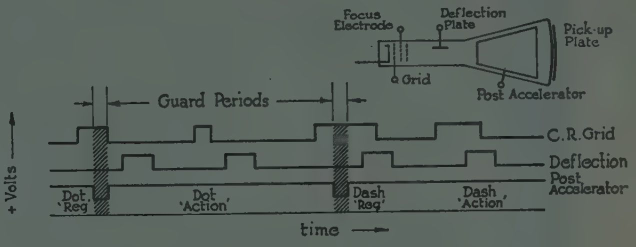
Focus Electrode
Deflection Plate
Pick-up Plate
Grid
Post Accelerator
Guard Periods
C.R.Grid
Deflection
Post Accelerator
+ Volts
Dot 'Reg'
Dot 'Action'
Dash 'Reg'
Dash 'Action'
time

Fig. 6—Pulsing the post accelerator electrode during the guard period.

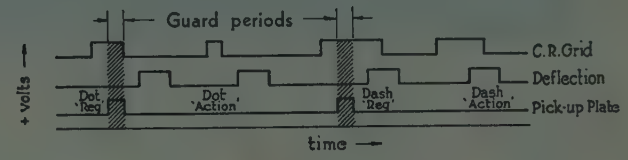
Guard periods
C.R.Grid
Deflection
Pick-up Plate
+ volts
Dot 'Reg'
Dot 'Action'
Dash 'Reg'
Dash 'Action'
time

Fig. 7—Pulsing the pick-up plate during the guard period.

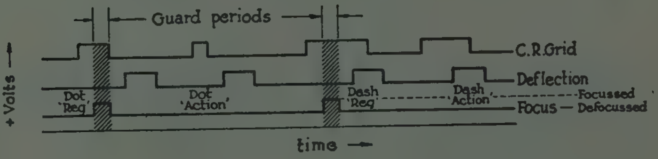
Guard periods
C.R.Grid
Deflection
Focussed
Focus — Defocussed
+ Volts
Dot 'Reg'
Dot 'Action'
Dash 'Reg'
Dash 'Action'
time

Fig. 8—Focusing the beam during the guard period.

The method shown in Fig. 5 has been adopted for the electronic computer model (UTEC) at the University of Toronto. This method appears to be most suitable because it results in the least amount of excessive signal at the input of the pick-up amplifier.

VI. Experimental Results

The method described above was tested on 2BP 1, 3 BP 1, 3 JP 1, and 3 JP 11 tubes. Modulation of the cathode-ray tube beam similar to the method illustrated in Fig. 5 was employed throughout these experiments. Fig. 9 is a photograph of the computer on which these experiments were carried out.

The cathode-ray tubes were connected as shown in Fig. 10. Note that the restoring circuit fixes the most negative potential of the bright-up pulse. Fig. 11 shows a photograph of typical bright-up pulses. Action and regeneration periods alternated.

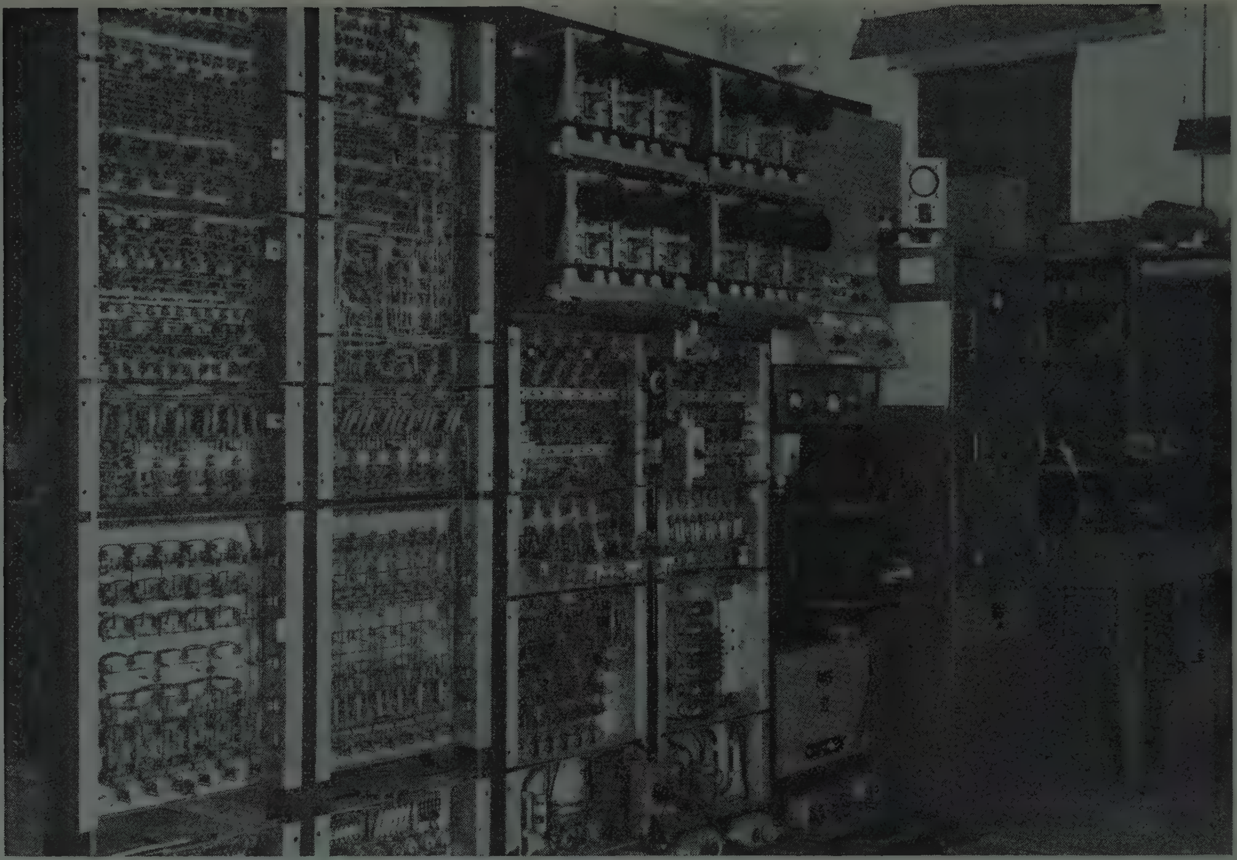

Fig. 9—University of Toronto electronic computer.

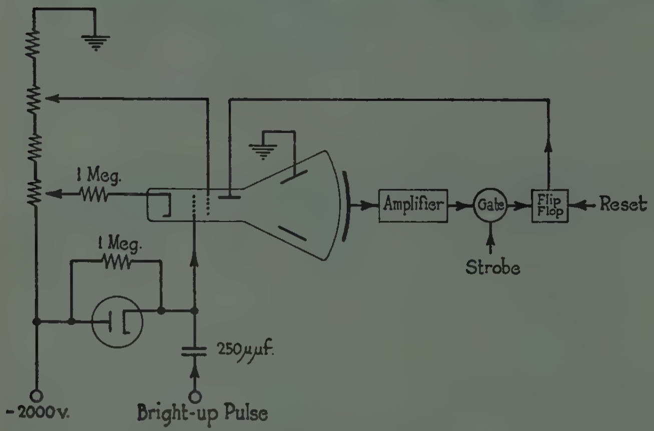

Fig. 10—Connection of cathode-ray tube.

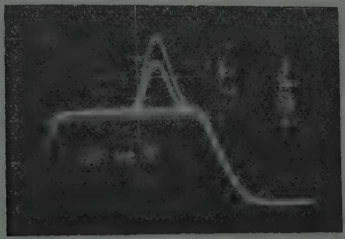

Fig. 11—Typical bright-up pulses.

During action periods, the bright-up pulse was always flat-topped and of about 3.5 μsecs duration. During regeneration periods, a voltage step E continuously variable from 0 to 40 volts could be superimposed on the flat-topped bright-up pulse. This voltage step was the only parameter changed during any one experiment. All other parameters were adjusted prior to the experiment. Spots storing positive charges are denoted "zeros" and spots storing negative charges are denoted "ones." Signals due to "ones" were adjusted prior to the experiment so as to be twice the amplitude necessary to just overcome the discrimination bias separating "ones" from "zeros."

Fig. 12 is a short exposure photograph of the face of the cathode-ray tube during the test. The spots are numbered in the following way:

```
      482          ......  510       512
 481       483     ......        511
      450          ......  478       480
 449       451     ......        479
 ...................................
 ...................................
 ...................................
 ...................................
      34           ......  62        64
 33        35      ......        63
      2         4  ......  30        32
 1         3    5  ......        31
```

During 2^n consecutive action periods ($n=2, 3, \cdots 9$) spot m is bombarded. During the next 2^n action periods spot $m+8$ is bombarded. After spot $512-p (p<8)$ has been bombarded the action beam returns to spot $9-p$ and so on. In this way every spot in the raster is subjected to 2^n consecutive action periods within a fraction of a second.

Initially all spots in the raster are cleared to "zeros." Due to secondary electron redistribution from the action spots those spots having a R.A.R. less than 2^n will change into "ones." Once a spot becomes a "one" it will remain a "one." The number of spots having a R.A.R. less than 2^n are counted in 30-second intervals. This number is plotted for various heights of the step E and for different values of the R.A.R. (2^n) in Figs. 13, 14 and 15, for three typical tubes. Results were similar for a total of 15 tubes tested. Results also were similar for the same tube when different operating conditions were employed, such as larger than optimum beam focus, smaller or larger raster size, different accelerating voltages and so forth. The actual R.A.R. would, of course, be sensitive to these variations in operating parameters, but the variation in R.A.R. with the step took the same form as illustrated in Figs. 13, 14, 15 and 16.

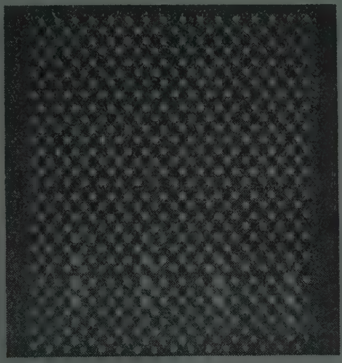

Fig. 12—Cathode-ray tube display during R.A.R. test.

A programmed R.A.R. test was also employed on a few tubes. In this test, four spots adjacent to a given spot S, (S being initially a "zero") are each bombarded for n successive action periods. S is then tested. If S is still a "zero" n is increased by unity and the above process repeats. This is continued until S changes into a "one" or until n reaches the value 128 which corresponds to a R.A.R. of 4 times 128, i.e. 512. The value of n is then printed out by the computer and a different spot S is selected. Fig. 16 shows typical values of the R.A.R. so obtained plotted against the height of the bright-up step E.

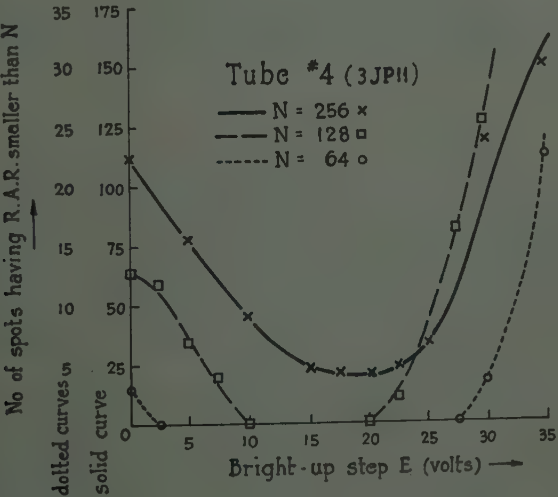

Fig. 13—Number of spots having R.A.R. less than N plotted against height of bright-up step E.

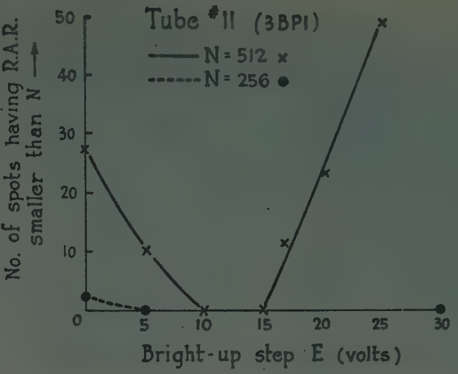

Fig. 14—Number of spots having R.A.R. less than N plotted against height of bright-up step E.

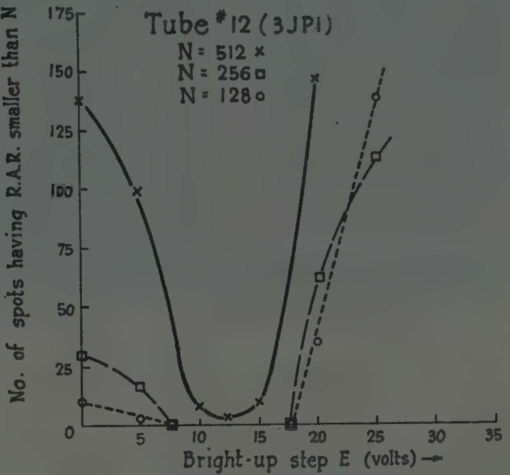

Fig. 15—Number of spots having R.A.R. less than N plotted against height of bright-up step E.

Figs. 13 to 16 show that the R.A.R. reaches a definite maximum and then decreases again as the bright-up step is increased. This would be expected from the following considerations.

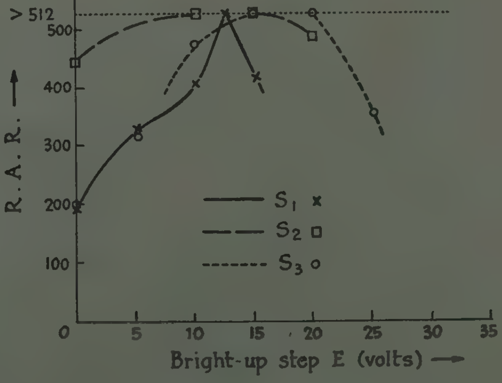

Fig. 16—Typical variation of R.A.R. with bright-up step for 3 spots, S_1, S_2, and S_3.

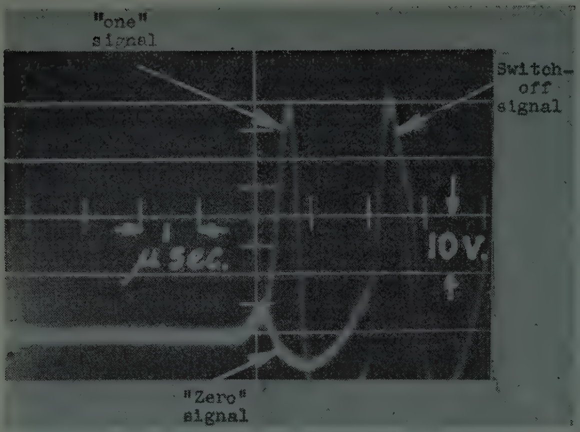

Fig. 17—Typical "one" and "zero" signals in the absence of interference.

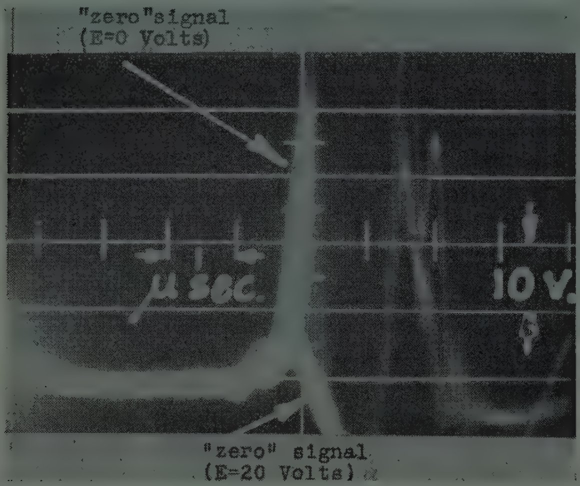

Fig. 18—Output signals from spot S adjacent to action-spot S_1 in the absence and presence of 20-volt bright-up step during guard period.

During each action bombardment of a spot adjacent to a given spot S, an average redistributed charge ρ_A will flow to S. During each regeneration bombardment of a spot adjacent to a given spot S, an average redistributed charge ρ_R will flow to S. Experiment shows that the effect of the eight spots immediately adjacent to S only needs to be considered. The total redistributed charge ρ which flows to S between two successive regenerations of S is given by

$$\rho = A\rho_A + 8\rho_R, \tag{1}$$

where A is the number of action bombardments of spots adjacent to S between two regenerations of S. The R.A.R. of S will vary inversely as ρ. Due to the extra bright-up step E during regenerations, ρ_R will be an increasing function of E and ρ_A will be a decreasing function of E. When E is zero ρ_A and ρ_R will be substantially equal. Since the maximum possible A is usually much greater than 8, the decrease in the first term in (1) will dominate the expression as E is increased from zero. ρ_A is bounded as E is increased since it cannot decrease below zero. ρ_R will increase with increasing E in an unbounded fashion.

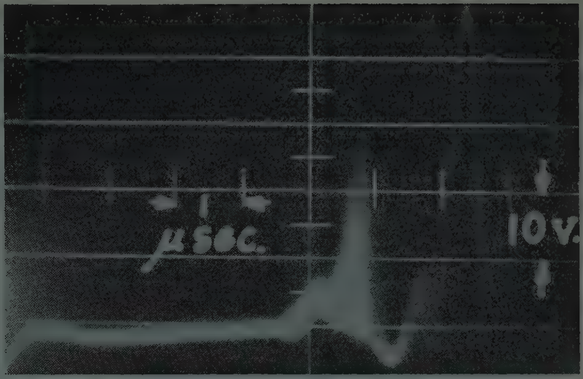

Fig. 19—Output signal from S due to 10-volt bright-up step during guard periods.

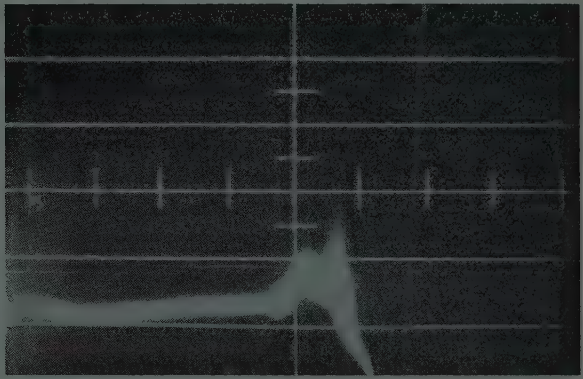

Fig. 20—Output signal from S due to 25-volt bright-up step during guard periods.

For example assume that ρ_A and ρ_R are exponentially dependent on E as given by

$$\rho_A = Ke^{-bE}; \qquad \rho_R = Ke^{bE}, \tag{2}$$

then

$$\rho(E) = AKe^{-bE} + 8Ke^{bE},$$

which is a concave upward curve having an initial value of

$$(A + 8)K,$$

and having a minimum value $\rho(E)$ min. $\equiv 2\ K\sqrt{A8}$ for a step E given by

$$E = \frac{1}{2b} \ln \frac{A}{8} \cdot$$

Little is known at present about the exact functional dependence of ρ on E. The optimum value of E must, therefore, be obtained by direct experiment. Experiment shows that the maximum R.A.R. as a function of E is broad enough so that a single adjustment for bright-up step height should suffice for a complete cathode-ray tube storage system.

Fig. 17 is a photograph of "zero" and "one" output signals in the absence of interfering secondary electron currents. Figs. 18, 19 and 20 are photographs of the signals from a spot S adjacent to a spot S_1. During every action period S_1 was bombarded. This resulted in the maximum flow of interfering secondary electron currents to S.

To prevent the "zero" stored at S from changing into a "one" the strobe pulse, shown in Fig. 10, was disconnected while the photographs of Figs. 18, 19, and 20 were taken. In Fig. 18 the output signals from S for no bright-up step and for a 20-volt bright-up step are superimposed. The output signal in the absence of a step would clearly cause the spot to change into a "one" if the regeneration loop were actuated by the strobe pulse. The lower signal shows how the step has reduced the interference and is almost as good as the "zero" signal of Fig. 17. Note that the step itself results in a negative output signal during the rise of the step and a positive signal during the fall of the step. The latter signal merges with the positive signal on beam switch off. As a result the switch-off signals increase with increasing steps. Fig. 19 shows the output signal from S for a 10-volt step, and Fig. 20 shows the output signal for a 25-volt step.

The blurring of the output signals from S is caused by changes in the positive interference signal during the time (1 min.) of the exposure. These changes are quite large and repetitive. They are believed to be due to the discontinuous nature of charge leakage on the insulator. It is believed that a wall of negative charge builds up between S and S_1. When the wall charges to a certain negative potential there is a sudden breakdown and some of the charges flow to adjacent portions of the screen resulting in a levelling of the screen potential in the neighbourhood of S. Just before this breakdown, the interfering currents to S will be at a minimum, and just after the breakdown, these currents will be at a maximum. The actual nature of the variations in the output signals from S strongly suggests such an explanation.

The variations are erratic yet repetitive. The positive interference signal always increases suddenly to a maximum and then decreases in a slower jumpy fashion to a minimum. The duration of the cycle itself varies quite greatly. This appears to rule out interference variations due to steady pulsations, such as raster repetition frequencies, power supply ripple and so on. Signals from other spots which receive a negligible flow of interfering currents were perfectly steady throughout the experiment.

The changes of the maximum interfering signal from an amplitude of 50 volts for no bright-up step, shown in Fig. 18, through an amplitude of about 30 volts for a 10-volt step, shown in Fig. 19, and an amplitude of about 6 volts for a 20-volt step, appearing in Fig. 18, to an amplitude of about 15 volts for a 25-volt step, shown in Fig. 20, are quite evident. The discrimination bias is 20 volts so that S would change into a "one" for steps up to 10 volts and would be able to store "zeros" for steps from 15 to 25 volts, in the presence of the interference from S_1.

VII. Conclusions

The method here proposed adds essentially a new function to the beam of cathode-ray storage tubes. This function is to create between storage locations a potential barrier which greatly reduces or prevents the flow of secondaries from action spots to adjacent storage locations. The potential barrier may be thought of as taking the place of the barrier grid in barrier-grid storage tubes. Thus the advantages of a barrier grid may be at least partially realized in a cathode-ray storage tube without actually incorporating such a grid in the tube.

This method may be combined with other methods[5] for improving the R.A.R. The benefits of the several methods should then be additive. Experimental work has shown that the method here described gives good results.

VIII. Acknowledgment

The writer's work was carried out under a grant of the Canadian Defence Research Board and the National Research Council of Canada. The writer also wishes to acknowledge with gratitude the technical assistance of L. Casciato and numerous discussions with colleagues at the University of Toronto and other projects.

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[5] R. Schumann, "Improvement of Williams' Memory Reliability," Paper presented at the Meeting of the A.C.M. at the University of Toronto; September 1952.

Stagger-Tuned Loop Antennas for Wide-Band Low-Frequency Reception*

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***Summary*—This paper describes a system of stagger-tuned high-*Q* loop antennas designed for 100 kc Loran pulse reception. The individual loop antenna outputs are first added in a summing amplifier followed by a grounded-grid amplifier and a cathode-follower stage before being fed to the receiver. Based upon preliminary theoretical analysis, an experimental antenna array is developed, which confirms its capability of achieving a wide bandwidth with relatively small dimensions. A squaring scheme of obtaining omnidirectional reception property is suggested; and a modified electrolytic tank method, which is adaptable to the present system in synthesizing a given response characteristic, is also described.**

INTRODUCTION

IN LOW-FREQUENCY LORAN TRANSMISSION, where the carrier frequency is in the neighborhood of 100 kc, efficiency and bandwidth considerations demand an antenna system of large dimensions if a vertical antenna is used. At 100 kc the wavelength is 3,000 meters; portable units cannot be provided with antennas of adequate length. If short antennas were used, the effective *Q* would be extremely high, and external damping must be provided in order to obtain the desired bandwidth for pulse reception. External damping reduces circuit efficiency. In the present system, high-*Q* loop antennas of small dimensions are used. The individual loop antennas are stagger-tuned to different frequencies within the required bandwidth. Their outputs are added in a summing amplifier, which is followed by a grounded-grid amplifier to improve stability and a cathode-follower stage to match the input impedance of the receiver unit. It proves to be a feasible and desirable substitute for long vertical antennas.

THE INPUT SUMMING AMPLIFIER

The input summing amplifier is used to add the signals received by the stagger-tuned loop antennas. The primary requirements on the part of the summing amplifier are: (1), small interaction among the different input channels, (2), low noise, (3), wide band, and (4), linear operation.

The arrangement chosen was a parallel-plate summing amplifier circuit. There were as many tube sections as the number of input channels. The plates of all tube sections were tied together and the output was taken from across a common plate tuned circuit properly loaded to give the required bandwidth. When identical tubes are used, the output voltage can readily be shown to be

$$e_0 = -\frac{g_m}{Y_L + nY_p} \sum_{i=1}^{n} e_{gi}, \tag{1}$$

where e_{gi} are input voltages at the grids; n, the total number of input channels; Y_L, the admittance of the common load circuit; and g_m and Y_p, the transconductance and the plate conductance respectively of the tubes. The weighting factors are the same for all channels and will decrease as the number of input channels increases. It is to be pointed out that the input voltages e_{gi} may be a quite complicated function of frequency and loop parameters. Type 6*J*6 miniature dual-triode tubes were chosen because of their high g_m (5,000 micromhos). Special care should be exercised in selecting the tubes and the associated circuit components to achieve uniformity for the different input channels.

If high-transconductance triodes like 6*J*6 are used, the noise-equivalent resistance is only several hundred ohms. This multi-tube parallel-plate arrangement has considerable improvement in the over-all noise figure of the receiving system as compared to a summing amplifier with resistive feedback and large isolating resistances in the grid circuit. It also possesses the advantages of high input impedance and almost zero coupling among input channels.

THE CASCODE AMPLIFIER AND CATHODE FOLLOWER STAGES

If the summing amplifier is connected directly to a sensitive receiver, the receiving system bursts into oscillation very easily when the gain of the receiver is set moderately high. In order to maintain the high gain of the receiver and achieve stability at the same time, a grounded-grid amplifier stage is placed between the summing amplifier and the input to the receiver. This arrangement of grounded-cathode (summing amplifier) and grounded-grid stages has been referred to as a cascode amplifier.[1] It combines the advantages of the high voltage amplification and stability of a pentode and the low noise factor of a triode. The input resistance of the grounded-grid stage is very low and approaches the reciprocal of the transconductance of the tube if its load resistance is small in comparison with the plate re-

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[1] H. Wallman, A. B. MacNee, and C. P. Gadsden, "A low-noise amplifier." PROC. I.R.E., vol. 36, p. 700; June 1948.

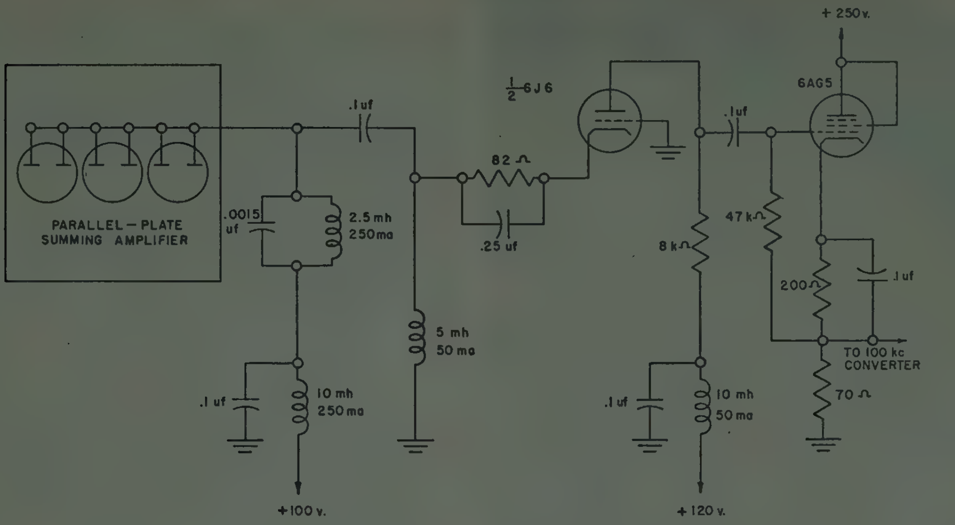

Fig. 1—Cascode amplifier and cathode follower stages.

sistance. This low resistance accounts for the low amplification of the summing amplifier stage and hence the greater stability. This arrangement also has a broad response with respect to frequency. With the circuit shown in Fig. 1, the 3-db band extends from about 40 to 200 kc, and the voltage gain over the range of 70 to 140 kc is substantially constant.

Lowering the load resistance in the grounded-grid stage would result in a lower input resistance and thus further improve stability and widen bandwidth, but it was found that this would accentuate the difference in the amplification of the different channels in the summing amplifier due to non-uniformity of tubes and other circuit elements. It would also reduce the over-all voltage amplification.

The input to the 100 kc converter leading to the Loran receiver has a series L-C tuned circuit for all four channels. The cascode amplifier arrangement of the grounded-cathode and grounded-grid stages as described above, however, has a high output impedance. In order to achieve a better impedance match, a cathode-follower stage is designed which serves as an impedance-transforming device. This circuit is included in Fig. 1.

ANALYTICAL DETERMINATION OF SUMMING AMPLIFIER OUTPUT WITH STAGGERED INPUTS

Before one can proceed to design a stagger-tuned loop antenna system, it is necessary to know the dependence of its resultant response upon such factors as the number of loops and their relative orientation, the spacing of the resonant frequencies, and the Q, the area, and the number of turns of each individual loop. One method of analyzing the resultant response would be to use the modified electrolytic tank method described in the Appendix. This method provides a convenient way of synthesizing the desired amplitude and phase responses on a cut-and-try basis. However, because of the lack of time for constructing and experimenting with an electrolytic tank, this method was not tried. Instead, an analytical method was developed which is described below and which may be used with relative ease to find the response characteristics of the stagger-tuned system.

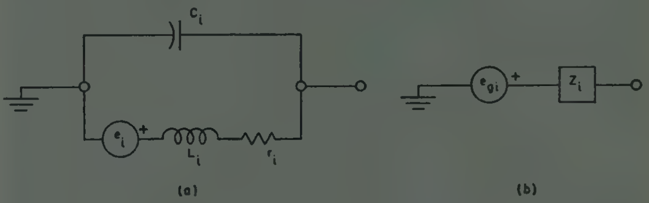

Fig. 2—Equivalent representations of tuned loop antenna circuit.

If each tuned loop antenna is schematically represented by the circuit in Fig. 2(a) and Fig. 2(b) as its series equivalent, then

$$e_{gi} = \frac{e_i}{\left(1 - \frac{\omega^2}{\omega_i^2}\right) + j\frac{1}{Q_i}\frac{\omega}{\omega_i}}, \tag{2}$$

$$Z_i = r_i\left[\frac{1 + jQ_i\frac{\omega}{\omega_i}}{\left(1 - \frac{\omega^2}{\omega_i^2}\right) + j\frac{1}{Q_i}\frac{\omega}{\omega_i}}\right], \tag{3}$$

where $\omega = 2\pi f$, f = frequency in cps, $\omega_i^2 = 1/L_iC_i$, and $Q_i = \omega_iL_i/r_i = Q$ of ith loop antenna at its resonant frequency. Substituting (2) and (1) and rearranging terms, one gets

$$e_0 = \frac{g_m}{Y_L + nY_p}\sum_{i=1}^{n}\frac{\omega_i^2 e_i}{2\beta_i^2} \cdot \left[\frac{1}{\left(\frac{\omega}{\beta_i} - 1\right) - j\frac{\alpha_i}{\beta_i}} - \frac{1}{\left(\frac{\omega}{\beta_i} + 1\right) - j\frac{\alpha_i}{\beta_i}}\right], \tag{4}$$

with

$$\alpha_i = \frac{r_i}{2L_i} = \frac{\omega_i}{2Q_i}, \qquad \beta_i = \omega_i\sqrt{1 - \frac{1}{4Q_i^2}},$$

and $\alpha_i^2+\beta_i^2=\omega_i^2$. Under the easily realizable condition $(1/4Q_i^2)\ll 1$, further simplification results because $\beta_i \doteq \omega_i$. Thus,

$$e_0 = \frac{g_m}{2(Y_L + nY_p)} \sum_{i=1}^{n} e_i \cdot \left[\frac{1}{\left(\frac{\omega}{\omega_i} - 1\right) - j\frac{1}{2Q_i}} - \frac{1}{\left(\frac{\omega}{\omega_i} + 1\right) - j\frac{1}{2Q_i}} \right]. \quad (5)$$

In the application under consideration, interest is centered on frequencies in the range of 90 to 110 kc. The ratio ω/ω_i then has values very nearly equal to 1. Since $1/2Q_i$ is usually very small compared with 2, the first terms in the square bracket, under the summation sign will in general be much larger than the corresponding second terms, each of the latter having a magnitude of 1/2 to a good approximation. If one substitutes $\Delta\omega_i$ for $\omega-\omega_i$, the typical first term can be written as

$$\frac{e_i}{\frac{\Delta\omega_i}{\omega_i} - j\frac{1}{2Q_i}} = (2Q_i e_i \sin\theta_i)\epsilon^{j\theta_i}, \quad (6)$$

with

$$\theta_i = \tan^{-1}\left(\frac{1}{2Q_i}\cdot\frac{\omega_i}{\Delta\omega_i}\right). \quad (7)$$

For loops of equal area, a change in the incoming angular frequency ω changes the angle θ_i; the relative values of the induced voltage e_i remain unchanged. The amplitude and phase response of the system can be computed with relative ease.

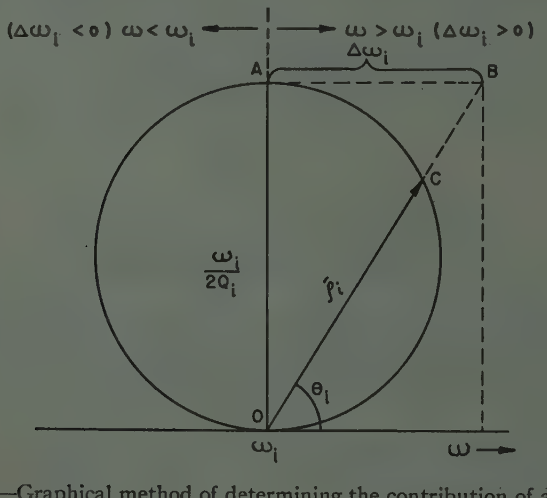

Fig. 3—Graphical method of determining the contribution of a loop.

Since one can treat (6) as a vector, a graphical method may also be conveniently employed. It is easy to show that if a circle of diameter $\omega_i/2Q_i=r_i/2L_i=\alpha_1$ is drawn on the frequency axis at ω_i for a given loop as shown in Fig. 3, then for any given angular frequency ω, which determines the length of $\overline{AB}=\Delta\omega_i$, the vector $\bar{\rho}_i$ represented by $\overline{OC}$ determines the relative contribution of this particular loop at the given frequency. It is noted that α_i is equal to one half of the 3-db bandwidth of the ith loop. The typical first term as expressed in (6) is $4(Q_i^2/\omega_i)e_i\bar{\rho}_i$. For the special case of identical loops, circles of equal diameter $r/2L$ may be drawn at the resonant angular frequencies ω_i of the individual loops, and the resultant response determined by adding the vectors $\bar{\rho}_i$ accordingly.[2]

The Loop Antennas

The effective height of a loop antenna for vertically polarized incoming wave is

$$h_e = \frac{2\pi NA}{\lambda}, \quad (8)$$

where h_e=effective height, N=total number of turns, A=area of the loop, and λ=wavelength of the incoming signal. Equation (8) holds true for loop antennas of any shape provided that the plane of the loop is in the direction of the incoming signal and that the thickness of loop sides is very small compared with loop dimensions. The effect of tuning and orientation on the resultant output is to be accounted for by a staggering or detuning factor, and a directional factor respectively.

Formulas are usually available for calculating the inductance and the resistance of a loop of given configuration and given number of turns when the size of the wire is known. But seldom is account taken of the effect of distributed capacitances. The effective inductance of a closely wound coil at 100 kc. may indeed be quite different from the so-called low-frequency inductance given by the formulas neglecting distributed capacitances.

Single-plane spiral-wound coils were used as the loops because of their inherent advantages of physical compactness, low distributed capacity and high-Q properties. The final design of the loops has the following physical data: Outside dimensions 20″×20″, Inside dimensions: 6″×6″, Total number of turns: 70, Size of wire: AWG No. 26, enamel covered. The measured effective inductance of the loop has an average value of 1.9 millihenrys and its Q_e is about 60 at 100 kc.

Experimental Investigation

Twelve identical loops were used in the experimental set-up. They were initially tuned to frequencies from 90 to 110 kc at 2 kc intervals, the twelfth loop being equipped with a variable air condenser which may be set at any desirable frequency within this range so as to smooth out the resultant amplitude response. There are an infinite number of ways to arrange the loop antennas. It must be settled at some compromise between the mutual coupling effect, the space occupied, and the reception characteristics. Data presented herein were obtained with all twelve loops in parallel. They were arranged such that no two loops would have mutual effects more than those arising from a space separation of 3 feet and a frequency separation of 4 kc.

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[2] For details and simplifying recurrence formulas, which have been developed for this purpose, the reader is referred to "Stagger-Tuned Loop Antennas for Wide-Band Low-Frequency Reception," Watson Laboratories Contract No. AF 28(099)-111, Final Report, Syracuse University· September 1950.

For taking the amplitude and phase characteristics of the antenna system as a function of frequency, a special wide-band trf receiver, consisting of two stages of single-tuned rf amplifier, two stages of resistance-capacitance coupled amplifier and a tuned power amplifier stage, was constructed. The over-all 3-db bandwidth of the receiver was over 20 kc and the amplification at the center of the band was over 4×10^6. There are two output positions on this receiver. At one position, the output from the power amplifier appears directly at the output terminals, which may be used to feed an indicator directly for cycle-matching purposes; at the other position, a detector stage is switched into the circuit but the by-pass condenser is purposely made small so that the number of rf cycles can still be counted on the rectified pulse envelope. The receiver was used in conjunction with the indicator section of a DAS-4 Loran set when the actual wave shape of the receiver pulses was photographed.

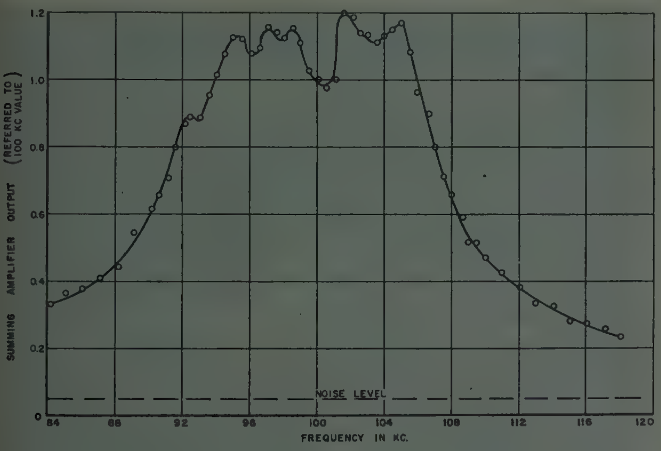

Fig. 4—Amplitude response of a 12-element stagger-tuned loop antenna system.

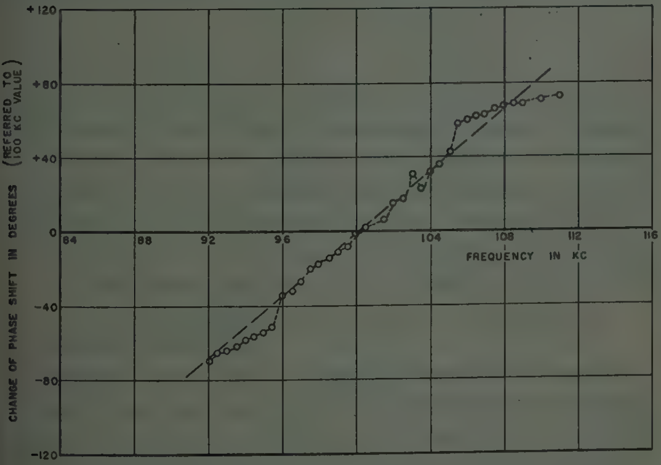

Fig. 5—Phase response of a 12-element stagger-tuned loop antenna system.

Figs. 4 and 5 show respectively the amplitude and phase response characteristics of the 12-element stagger-tuned loop antenna system. It is seen that the 3-db bandwidth of this system is about 16.5 kc and that the amplitude response outside of this frequency band drops quite rapidly, thus reducing noise pick-up outside of the band where no significant signal frequency components are present. The phase characteristic shows the existence of a linear relationship between the phase shift and the frequency within the "pass-band." This linear phase characteristic is essential for pulse reproduction. There are minor peaks and dips in the amplitude response, but they do not prove to be important. The number of peaks does not correspond to the number of loop antennas because some of the peaks partly merge and the presence of mutual-coupling effects has a tendency to make this more so. Through meticulous retuning, the response curves can doubtlessly be made more smooth.

The shapes of the Loran pulses as received through the trf receiver and DAS-4 indicator combination using stagger-tuned loop antennas under different conditions were recorded photographically. Fig. 6(a) gives the shape of the pulse at the transmitter which was located 40 miles away from the receiving site, and Fig. 6(b) shows the envelope of the received pulse when all twelve loops were used and properly tuned. Each rf cycle corresponds to a duration of 10 microseconds

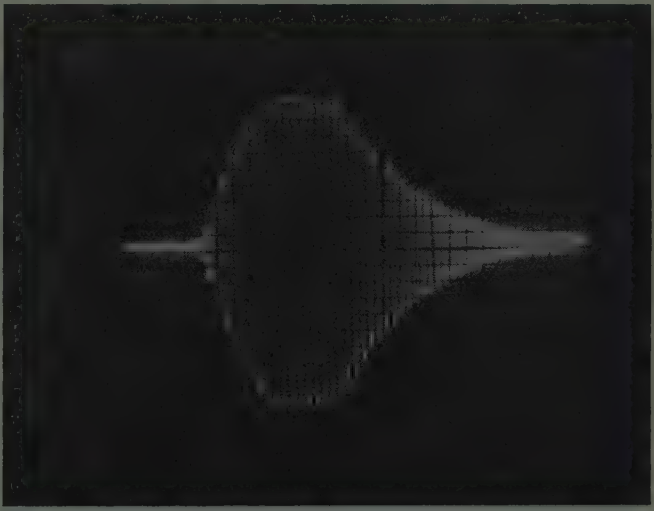

Fig. 6(a)—Pulse shape at the transmitter.

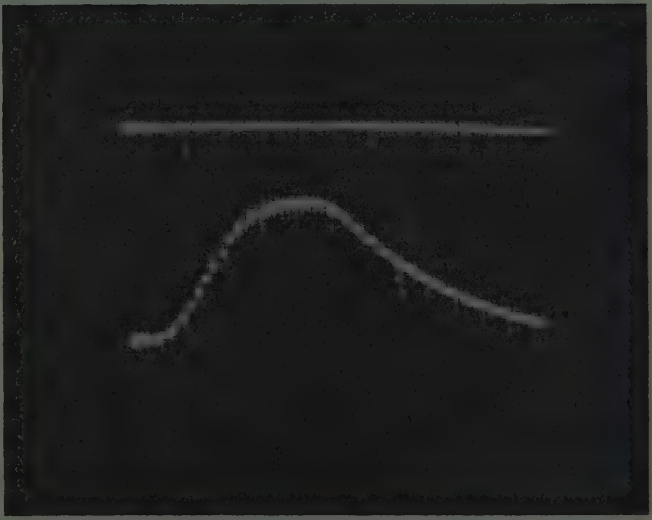

Fig. 6(b)—Envelope of received pulse with twelve stagger-tuned loops.

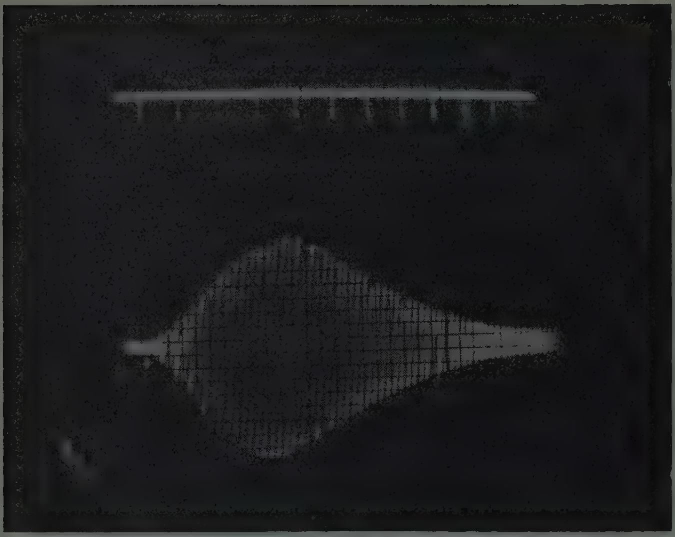

Fig. 7(a)—Received pulse with two lf and two hf loops disconnected.

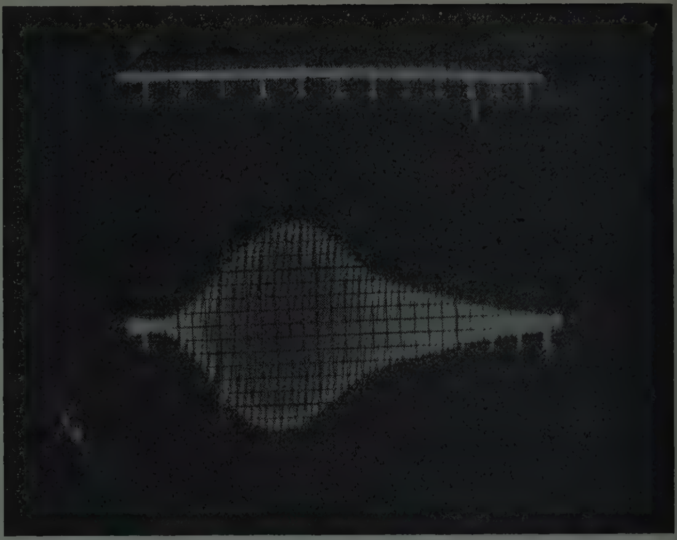

Fig. 8(a)—Received pulse with three lf loops disconnected.

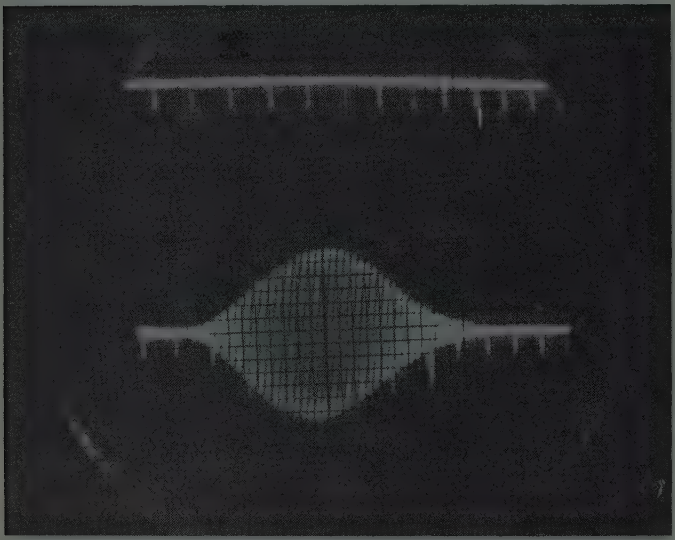

Fig. 7(b)—Received pulse with three lf and three hf loops disconnected.

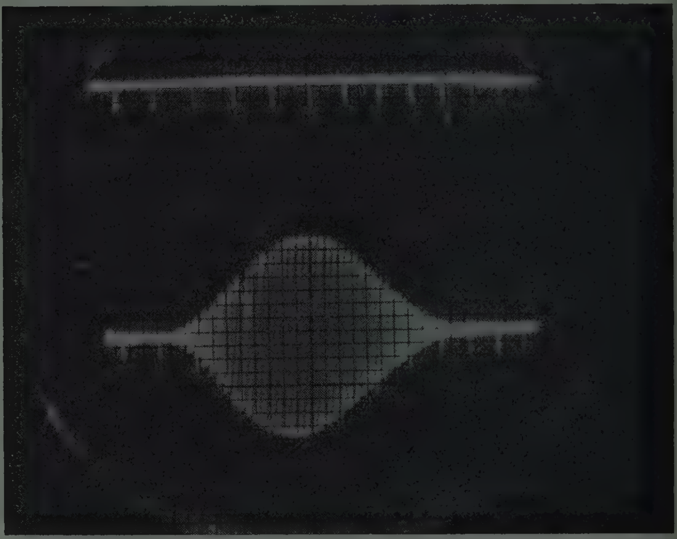

Fig. 8(b)—Received pulse with three hf loops disconnected.

at 100 kc. By comparing these two figures, it is seen that the received pulse represents a quite good reproduction of the pulse at the transmitter. The rise time is somewhat lengthened, but there is no serious deterioration of the pulse shape.

Fig. 7(a) shows the shape of the received pulse when two loops from both the low-frequency and the high-frequency ends of the assigned band were disconnected from the summing amplifier, leaving the tuning and the position of the other eight loops unchanged. It is seen that the rise time is further lengthened and that, the flat portion of the pulse is no longer visible, although the amplitude of the pulse is not much reduced. Both of these effects are accentuated in Fig. 7(b) when three loops are taken out from both ends of the assigned band. Here, the reduction in the amplitude of the pulse is more obvious. The removal of end loops reduces both the amplitude and the effective bandwidth of the resultant amplitude response curve.

The shapes of the received pulses with three low-frequency loops disconnected, and with three high-frequency loops disconnected are shown respectively in Figs. 8(a) and 8(b). The removal of the lf loops resulted in a prolonged delay time. It is evident that Fig. 6(b), with all twelve loops in the circuit and properly tuned, represents by far the best response.

Scheme of Achieving Omnidirectivity

For a wide-band antenna system to be omnidirectional to pulse signals, it should be able to give the same response to signals of frequencies spread over the entire band from all directions. If the response characteristic varies appreciably with signal direction, the system will not be suitable for applications like Loran, wherein pulses from master and slave stations are to be accurately compared.

Arranging a large number of loops radially or in the form of a rectangular polygon does not present a solution. Fig. 9 depicts the squaring scheme of obtaining omnidirectional reception property from a loop antenna system.

As indicated in the figure, the stagger-tuned loops are arranged in two identical groups placed at right angles to each other and the outputs of both groups are passed through a squaring circuit and then added in a summing amplifier. Ignoring the change in magnitude while undergoing squaring and summing operations, which can easily be compensated for, the resultant output is the square of the original signal; the directional angle θ does not influence the response. Although nonlinear operation is involved in the process, it is not objectionable because one is interested primarily in the comparison of pulse signals coming from different directions and not in the faithful reproduction of any signal carrying intelligence. Fig. 9 is only schematic, and the loops should not be placed too closely together in practice.

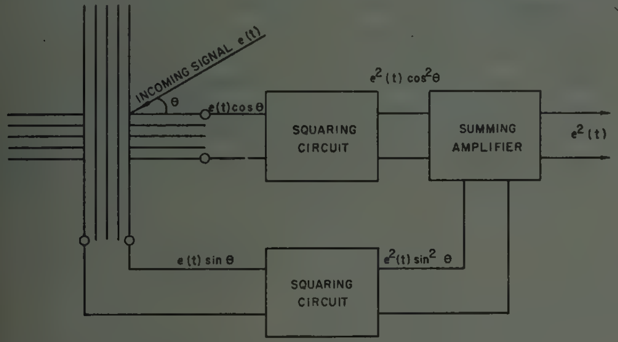

Fig. 9—The squaring scheme for achieving omnidirectivity.

CONCLUSIONS

The present 12-element stagger-tuned loop antenna system has a 3-db bandwidth of 16.5 kc and a phase shift which varies very nearly linearly with frequency over the desired range. The effective height of the antenna system alone was found to be about 3 meters, and the gain of the parallel-plate summing amplifier with 12 input channels is in the neighborhood of 4. These properties together with the experimental evidence of Loran pulse reception enable one to conclude that a system of high-Q loop antennas of small dimensions properly tuned to different frequencies within the required frequency band is a feasible and desirable substitute of long vertical antennas for wide-band low-frequency reception. The squaring scheme provides the means of achieving of omnidirectivity.

Generally, an effective height of 3 meters is not required. With the existence of sensitive receivers, an effective height of one half meter may be sufficient. In that case, the loops can be made with smaller dimensions and fewer turns. The effective Q of the individual loops would then be lower and it would be easier to tune the system to give a uniform amplitude response. However, analyses[3] show that the noise figure of a tuned loop antenna is very nearly in direct proportion to the loss resistance of the loop and is inversely proportional to its radiation resistance. The use of loops of lower Q would then result in lower circuit efficiency and less improvement in the noise figure and the absolute sensitivity[4] of the system.

It is possible that loops of fewer turns and smaller dimensions can be made in such a way that the effective Q is not lowered. The feasibility of employing iron-cored loops should also be well worth investigating.

ACKNOWLEDGMENT

The authors wish to acknowledge their indebtedness to Messrs. R. N. Lothes and H. G. Pascalar for their valuable contributions in this work, and to Captain O. D. Goldberg of Rome Air Development Center, Rome, New York, for profitable discussions on this project.

APPENDIX

Modified Electrolytic Tank Method of Analyzing the Response of a Stagger-Tuned Loop Antenna System

A rational function of a complex variable $s=\sigma+j\omega$ can be written in the following form:

$$F(s) = K\frac{\prod_{j=1}^{n_z}(s - s_{zj})}{\prod_{k=1}^{n_p}(s - s_{pk})}, \tag{A-1}$$

where K is a constant, and where the continued product in the numerator indicates that the function has n_z zeros, and that in the denominator shows that it has n_p poles. If the logarithm of (A-1) is taken, it will be seen that the logarithm of the magnitude and the phase of $F(s)$ are analogous to the expression for the electrostatic potential distribution around n_z equal negative line charges placed at positions corresponding to the zeros and n_p similar positive line charges at the locations of the poles. Consequently, with a suitable choice of constants, equipotential curves around these line charges correspond to curves of constant absolute magnitude of $F(s)$, and an orthogonal set of curves corresponds to lines of constant phase of $F(s)$.

In homogeneous and isotropic medium where the Ohm's law is obeyed, the results of electrostatic theory may be applied directly to the state of dynamic equilibrium represented by a steady current flow. Hansen and Lundstrom[5] have utilized this principle to determine experimentally the magnitude and phase of impedance and gain functions by means of an electrolytic tank. It should be stressed, however, that the usefulness of the electrolytic tank method depends primarily upon the determination of the positions in the complex plane of the zeros and poles of the function concerned. This method will, therefore, be ineffective if the zeros and/or poles of the functions under consideration are difficult to locate.

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[3] H. G. Pascalar, Master's thesis, L. C. Smith College of Engineering, Syracuse University; October 1950.

[4] D. O. North, "The absolute sensitivity of radio receivers," *RCA Review*, vol. 16, p. 337; January 1942.

[5] W. W. Hansen and O. C. Lundstrom, "Experimental determination of impedance functions by the use of an electrolytic tank," PROC. I.R.E., vol. 33, pp. 528–534; August 1945.

From (1) and (2) in the text and replacing $j\omega$ by s, one can easily show that

$$F(s) = e_0(s) = -\frac{g_m}{Y_L + nY_p}\sum_{i=1}^{n}\frac{\omega_i^2 e_i}{(s - s_i)(s - s_i^*)}, \qquad \text{(A-2)}$$

where the conjugate poles are

$$s_i = -\frac{r_i}{2L_i} + j\sqrt{\frac{1}{L_iC_i} - \frac{r_i^2}{4L_i^2}} = -\alpha_i + j\beta_i, \qquad \text{(A-3)}$$

$$s_i^* = -\frac{r_i}{2L_i} - j\sqrt{\frac{1}{L_iC_i} - \frac{r_i^2}{4L_i^2}} = -\alpha_i - j\beta_i. \qquad \text{(A-4)}$$

For a single loop antenna ($n=1$) there will be two poles at s_1 and s_1^* in the complex plane and two zeros at infinity. For two loops ($n=2$), two additional poles at s_2 and s_2^* and two additional zeros $\bar{s}_a$ and $\bar{s}_a^*$ result, where the zeros $\bar{s}_a$ and $\bar{s}_a^*$ have to be determined from the following quadratic equation:

$$\omega_1^2(s - s_2)(s - s_2^*) + \omega_2^2(s - s_1)(s - s_1^*) = 0.$$

With n loops, an algebraic equation of $2(n-1)$th degree has to be solved in order to determine the new zeros. This may prove to be very difficult and render the electrolytic tank method impractical. The modified method[6] to be described below solves this difficulty.

Expressing (A-2) in partial fraction form:

$$F(s) = -\frac{g_m}{Y_L + nY_p}\sum_{i=1}^{n}\frac{\omega_i^2 e_i}{j2\beta_i}\left[\frac{1}{s + \alpha_i - j\beta_i} - \frac{1}{s + \alpha_i + j\beta_i}\right]. \qquad \text{(A-5)}$$

But $s = \sigma + j\omega$, hence

$$F(s) = \frac{g_m}{Y_L + nY_p}\sum_{i=1}^{n}\frac{\omega_i^2 e_i}{2\beta_i} \times\left[\frac{(\omega - \beta_i) + j(\sigma + \alpha_i)}{(\sigma + \alpha_i)^2 + (\omega - \beta_i)^2} - \frac{(\omega + \beta_i) + j(\sigma + \alpha_i)}{(\sigma + \alpha_i)^2 + (\omega + \beta_i)^2}\right]. \qquad \text{(A-6)}$$

Now consider the case of a *negative* line charge of strength D_k per unit length located at each pole in the *second quadrant* and an equal *positive* line charge at each pole in the *third quadrant*. The potential due to these line charges at any point $s = \sigma + j\omega$ in the complex plane will be

$$V = \frac{1}{B}\sum_{k=1}^{n} D_k\left[\ln\sqrt{(\sigma + \alpha)^2 + (\omega - \beta_k)^2} - \ln\sqrt{(\sigma + \alpha)^2 + (\omega + \beta_k)^2}\right], \qquad \text{(A-7)}$$

where B is a constant whose value depends on the permittivity of the medium and the units employed. From (A-7), one easily obtains,

$$\frac{\partial V}{\partial \sigma} = \frac{1}{B}\sum_{k=1}^{n} D_k \times\left[\frac{\sigma + \alpha}{(\sigma + \alpha)^2 + (\omega - \beta_k)^2} - \frac{\sigma + \alpha}{(\sigma + \alpha)^2 + (\omega + \beta_k)^2}\right] \qquad \text{(A-8)}$$

$$\frac{\partial V}{\partial \omega} = \frac{1}{B}\sum_{k=1}^{n} D_k \times\left[\frac{\omega - \beta_k}{(\sigma + \alpha)^2 + (\omega - \beta_k)^2} - \frac{\omega + \beta_k}{(\sigma + \alpha)^2 + (\omega + \beta_k)^2}\right], \qquad \text{(A-9)}$$

and

$$\frac{\partial V}{\partial \omega} + j\frac{\partial V}{\partial \sigma} = \frac{1}{B}\sum_{k=1}^{n} D_k \times\left[\frac{(\omega - \beta_k) + j(\sigma + \alpha)}{(\sigma + \alpha)^2 + (\omega - \beta_k)^2} - \frac{(\omega + \beta_k) + j(\sigma + \alpha)}{(\sigma + \alpha)^2 + (\omega + \beta_k)^2}\right]. \qquad \text{(A-10)}$$

Comparing (A-6) and (A-10), it is readily seen that if as many pairs of electrodes as there are pairs of conjugate poles (which equals the number of loop antennas) are put in the electrolytic tank, negative in the second quadrant and positive in the third quadrant, the potential gradient in the ω and σ directions along the real frequency axis will be proportional respectively to the real and imaginary parts of the response function $F(s)$ at real frequencies. One condition must here be satisfied, that is:

$$\left(\frac{\omega_i^2 e_i}{\beta_i D_i}\right) = \text{Constant} \qquad \text{(A-11)}$$

i = any number from 1 to n, or

$$D_i = \frac{\omega_i^2 e_i}{\omega_1^2 e_1}\cdot\frac{\beta_1}{\beta_i}D_1. \qquad \text{(A-12)}$$

When dissipation is low ($\alpha_i \ll \beta_i$, $\omega_i \doteq \beta_i$) and if the loops are identical and with the same orientation ($e_i = e_1$), one can write

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[6] Dr. W. R. LePage, Professor of Electrical Engineering, Syracuse University.

$$D_i \doteq \frac{\omega_i}{\omega_1} D_1. \tag{A-13}$$

In practice, (A-13) means that the current flowing to or from an electrode should be approximately proportional to its distance from the origin in the complex plane.

This modified method has the advantage that only the positions of the poles need be known. There is no need of solving algebraic equations of high order because the location of zeros does not have to be determined. The effect on the resultant response function by the addition of an extra loop-antenna input channel can be determined experimentally by introducing another pair of electrodes without additional difficulty.

Amplitude Stability in Oscillating Systems*

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Summary—As a supplement to the Kryloff and Bogoliuboff method of determining amplitude of oscillation in quasilinear systems, a method based upon energy balance is presented. From the energy balance condition, a criterion for stability of oscillation is deduced for the case of one degree of freedom. The treatment is then generalized to systems of n degrees of freedom, and the n stability criteria are derived in terms of rates of change, with respect to amplitude, of the power supplied and dissipated in the n oscillations.

MUCH INSIGHT INTO the problem of amplitude stability in oscillating systems can be gained by a consideration of the general oscillatory circuit of Fig. 1.

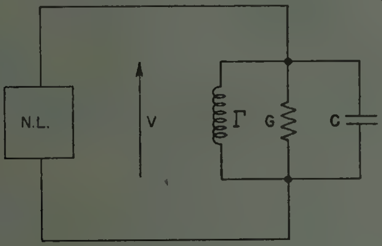

Fig. 1—Equivalent circuit.

In this circuit N. L. represents a nonlinear element which performs the dual functions of: (1) supplying the energy to overcome losses in G, the tank circuit shunt conductance, and (2) limiting the amplitude of oscillation to some finite value. The amount of nonlinearity need not be large. In fact, the analysis presented here applies only to systems in which the nonlinearity is small. However, virtually all sine wave oscillators belong to this class.

The volt-amp. characteristic of the nonlinear element may be represented by:

$$i = \alpha v + \beta v^2 + \gamma v^3 + \epsilon v^4 + \rho v^5, \tag{1}$$

and G_n, the nonlinear conductance is

$$G_n = \frac{di}{dv} = \alpha + 2\beta v + 3\gamma v^2 + 4\epsilon v^3 + 5\rho v^4. \tag{2}$$

The differential equation for the circuit is:

$$C\ddot{v} + (G + G_n)\dot{v} + \Gamma v = 0 \tag{3}$$

If the nonlinearity is small, this equation may be written as:

$$C\ddot{v} + \Gamma v + \mu f(v, \dot{v}) = 0 \tag{4}$$

where μ is a small parameter. The method of Kryloff and Bogoliuboff[1] may be applied to give as the solution:

$$v = a\epsilon^{-(G+G_e)t/2C} \cos \omega t \tag{5}$$

where:

$$G_e = \frac{1}{\pi a}\int_0^{2\pi} f(a \cos \theta) \cos \theta d\theta. \tag{6}$$

This equivalent linearized conductance is one which gives rise to the same fundamental component of current as the actual nonlinear conductance when a voltage of $v = a \cos \omega t$ is impressed.

For a nonlinear element such as that of (1), we obtain:

$$G_e = \frac{1}{\pi a}\int_0^{2\pi} [\alpha a \cos \theta + \beta a^2 \cos^2 \theta + \gamma a^3 \cos^3 \theta] \cos \theta d\theta \tag{7}$$

$$= \alpha + \tfrac{3}{4}\gamma a^2, \tag{8}$$

where we have used only 3 terms of the polynomial for simplicity. At equilibrium:

$$G + G_e = 0$$

$$a^2 = -\frac{4}{3\gamma}(\alpha + G). \tag{9}$$

An alternate method of solving for the equilibrium amplitude of oscillation consists of recognizing that the steady-state amplitude is that value at which the average energy supplied by the nonlinear element is equal to the average energy dissipated in the linear passive part of the system.

$$-\lim_{T\to\infty}\left[\frac{1}{T}\int_0^T ivdt\right] = \frac{a^2 G}{2}. \tag{10}$$

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[1] N. Minorsky, "Introduction to Nonlinear Mechanics," Chap. 10. J. W. Edwards, Ann Arbor, Michigan; 1947.

If the nonlinearity is sufficiently well represented by a third-degree polynomial, then this leads to:

$$a^2 = -\frac{4}{3\gamma}(\alpha + G) \qquad (11)$$

which is the same value we found in (9). This result has also been derived by Van der Pol.[2]

Graphical representations of power supplied by the nonlinear element and dissipated by the linear element may be used to determine stable and unstable levels of amplitude. P_D, the power dissipated, is a parabolic function of amplitude, and P_S, the power supplied, may be any shape curve, depending on the nonlinearity. Fig. 2 shows some possible curves.

Each point of intersection of a P_D curve with a P_S curve is a potential stationary value for the amplitude of oscillation. However, each such point must be inspected for stability. We may imagine the amplitude to be perturbed from its stationary value and determine the effect this perturbation has on the exponential factor in (5) which determines the growth or decay of the oscillation. If the sign of the exponent is such as to return the amplitude to its value prior to the perturbation, then the equilibrium is a stable one. However, if the sign of the exponent is such as to increase the change in the amplitude in the direction of the perturbation, then the equilibrium is unstable. This is equivalent to stating that only those points of intersection are stable for which

$$\frac{\partial P_D}{\partial a} > \frac{\partial P_S}{\partial a}. \qquad (12)$$

We must also recognize that there may exist semistable points of equilibrium, which are stable or unstable for either an upward or a downward perturbation of the amplitude, but not for both.

We may now identify as stable points in Fig. 2 the intersection in 2(a), points 1 and 3 in 2(c), point 2 in 2(d), point 2 in 2(e), and 3 in 2(f). Unstable points are the intersection in 2(b), point 2 in 2(c), point 1 and 2(d), point 3 in 2(e), and 2 in 2(f). Semistable points are the points of tangency in 2(e) and 2(f). Thus point 1 of 2(e) is stable for a downward perturbation of the amplitude and unstable for an upward perturbation, while the converse is true of point 1 in 2(f).

Since it is not possible in a physical system for the energy supplied or dissipated to be a multiple valued function of the amplitude, it is evident that stable and unstable points of equilibrium must alternate. However, it is also necessary to include semistable points in this generalization. In the language of nonlinear mechanics, stable and unstable limit cycles must alternate, and a limit cycle stable from the inside but unstable from the outside must enclose an unstable limit cycle and be enclosed by a stable limit cycle, while the converse is true for a limit cycle unstable from the inside but stable from the outside.

The possibility of simultaneous oscillations occurring at two or more different frequencies in the same circuit is well-known and has frequently been described and investigated.[3–5] The problem of determining the stability of such multiple oscillations may be attacked by methods similar to those we have employed here for circuits of one degree of freedom. Thus, for two degrees of freedom power supplied and power dissipated become functions of a_1 and a_2, the two amplitudes of oscillation.

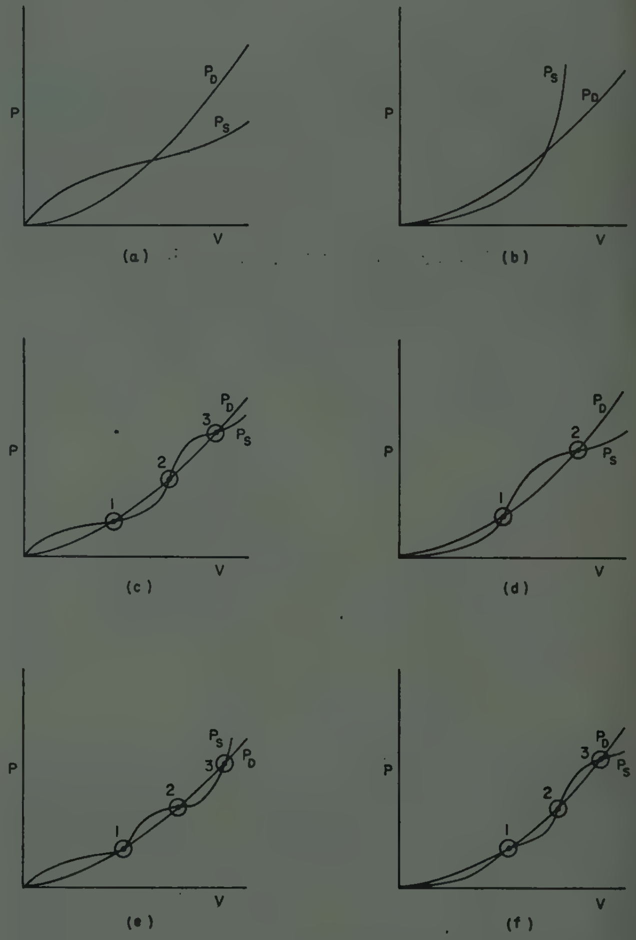

Fig. 2—Curves of power supplied and power dissipated vs. amplitude of voltage.

Although for one degree of freedom a simple criterion regarding the relative slopes of P_S and P_D at a point of equilibrium was sufficient to determine the stability, one intuitively expects the situation to be more complex for two degrees of freedom. Thus along a line of intersection of P_S and P_D there is an infinite number of points characterized by the same relative slopes of P_S and P_D.

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[2] B. Van der Pol, "The nonlinear theory of electric oscillations," Proc. I.R.E., vol. 22, pp. 1051–1086; September, 1934.

[3] L. V. Skinner, "Criteria for Stability in Circuits Containing Nonlinear Resistance," Ph.D. Thesis, University of Illinois; June, 1948.

[4] W. H. Huggins, "Multifrequency bunching in reflex klystrons," Proc. I.R.E., vol. 36, pp. 624–630; May, 1948.

[5] J. S. Schaffner, "Almost Sinusoidal Oscillations in Nonlinear Systems," University of Illinois Engineering Experiment Station, Bulletin Series No. 395; May, 1951.

Some additional criterion is needed to determine which, if any, of these points is stable. To do this we employ the method of perturbations, taking the initial perturbations to be small enough so that the calculation of their growth or decay may be handled as a linear problem. We shall derive the criteria for n degrees of freedom and then apply these results to the cases of two and three degrees of freedom. We define:

P_{Sj} = power supplied to the jth oscillation
P_{Dj} = power dissipated in the jth oscillation
a_j = amplitude of the jth oscillation

If all a_j undergo small perturbations, all P_{Sj} will also be perturbed. Evidently:

$$\frac{dP_{Sj}}{dt} = \sum_{i=1}^{n} \frac{\partial P_{Sj}}{\partial a_i} \cdot \frac{da_i}{dt} = \sum_{i=1}^{n} \frac{\partial P_{Sj}}{\partial (a_i^2)} \cdot \frac{d(a_i^2)}{dt} \cdot \tag{13}$$

If the system is stable, all P_{Sj} and a_j will return to their initial values, and we will have

$$\frac{dP_{Sj}}{dt} \longrightarrow 0,$$

which will occur when all

$$\frac{d(a_i^2)}{dt} \longrightarrow 0.$$

It may be shown that:

$$\frac{d(a_i^2)}{dt} = -\frac{2a_i^2\delta_{0i}}{G_i}(G_i + G_{ei}), \tag{14}$$

where δ_{0i} is the logarithmic decrement of the linear circuit at the frequency Ω_i of the ith oscillation, G_i is the conductance of the linear circuit at Ω_i, and G_{ei} is the equivalent linearized conductance of the nonlinear circuit at Ω_i.

Now let all amplitudes of oscillation undergo small perturbations:

$$a_i'^2 = a_{is}^2 + \Delta_i, \tag{15}$$

where a_{is} is the steady-state amplitude at the point where we are investigating stability, a_i' is the perturbed amplitude, and Δ_i is the difference of their squares. Then:

$$\frac{d(a_i'^2)}{dt} = \frac{d(a_{is}^2)}{dt} + \frac{d\Delta_i}{dt}$$
$$= -\frac{2(a_{is}^2 + \Delta_i)\delta_{0i}}{G_i}\left[G_i + f_i(a_{1s}^2 + \Delta_1, a_{2s}^2 + \Delta_2, \cdots, a_{ns}^2 + \Delta_n)\right], \tag{16}$$

where $f_i(a_{is}^2+\Delta_1, a_{2s}^2+\Delta_2, \cdots a_{ns}^2+\Delta_n)$ is the equivalent linearized conductance at frequency Ω_i and amplitudes $a_1 \cdots a_n$. Of course, the term $d(a_{is}^2)/dt$ is zero, since a_{is} is the steady-state amplitude. However, it will be convenient to carry the term along in our equations.

We have now:

$$\frac{d\Delta_i}{dt} = \frac{d(a_i'^2)}{dt} - \frac{d(a_{is}^2)}{dt}$$
$$= -\frac{2a_{is}^2\delta_{0i}}{G_i}\left[f_i(a_{1s}^2 + \Delta_1, \cdots, a_{ns}^2 + \Delta_n) - f_i(a_{1s}^2, \cdots, a_{ns}^2)\right]$$
$$- \frac{2\Delta_i\delta_{0i}}{G_i}\left[G_i + f_i(a_{1s}^2 + \Delta_1, \cdots, a_{ns}^2 + \Delta_n)\right]. \tag{17}$$

Now a Taylor's Series expansion of f_i can be made and substituted into this equation. However, since we are postulating small perturbations of the system, we may omit second and higher order terms. We thus obtain:

$$\frac{d\Delta_i}{dt} = -\left[\frac{2a_{is}^2\delta_{0i}}{G_i}\sum_{k=1}^{n}\frac{\partial f_i}{\partial(a_k^2)}\Delta_k\right] - \frac{2\Delta_i\delta_{0i}}{G_i}\left[G_i + f_i(a_{1s}^2, \cdots, a_{ns}^2)\right], \tag{18}$$

which is a system of n first order differential equations each containing n unknowns. Equation (18) may be rewritten as:

$$\dot{\Delta}_i = \sum_{j=1}^{n} a_{ij}\Delta_j \tag{19}$$

where

$$a_{ii} = \frac{-2\delta_{0i}}{G_i}\left[a_{is}^2\frac{\partial f_i}{\partial(a_i^2)} + G_i + f_i(a_{1s}^2, \cdots, a_{ns}^2)\right] \tag{20}$$

$$a_{ij} = -\frac{2a_{is}^2\delta_{0i}}{G_i} \cdot \frac{\partial f_i}{\partial(a_j^2)} \cdot \tag{21}$$

The characteristic equation for this system is:

$$\begin{vmatrix} (a_{11} - \rho) & a_{12} & \cdots a_{1n} \\ a_{21} & (a_{22} - \rho) & \cdots \\ \vdots & & \\ a_{n1} & & \cdots (a_{nn} - \rho) \end{vmatrix} = 0. \tag{22}$$

The roots of this equation are the exponents in the solutions for the Δ_i, and evidently must have no positive parts if the perturbations are to die out, i.e., if the amplitudes of oscillation are to be stable. This condition is insured by Routh's criterion, which, in the case of three degrees of freedom, requires that:

(a) $a_{11} + a_{22} + a_{33} < 0,$

(b) $-(a_{11} + a_{22} + a_{33}) \cdot (a_{11}a_{22} + a_{11}a_{33} + a_{22}a_{33} - a_{12}a_{21} - a_{13}a_{31} - a_{23}a_{32}) - (a_{11}a_{22}a_{33} - a_{11}a_{23}a_{32} - a_{12}a_{21}a_{33} + a_{12}a_{31}a_{33} + a_{13}a_{21}a_{32} - a_{13}a_{31}a_{22}) > 0,$ (23)

(c) $a_{11}a_{22}a_{33} - a_{11}a_{32}a_{23} - a_{12}a_{21}a_{33} + a_{12}a_{31}a_{23} + a_{13}a_{21}a_{32} - a_{13}a_{31}a_{22} > 0.$

For two degrees of freedom the criteria become:

$$\text{(a)} \qquad a_{11} + a_{22} < 0,$$

$$\text{(b)} \qquad -(a_{11} + a_{22}) \cdot (a_{11}a_{22} - a_{12}a_{21}) > 0, \qquad (24)$$

which may be obtained from (23) by eliminating all terms having 3's in the subscripts.

The power supplied at each frequency can be expressed in terms of the amplitude of that oscillation and the equivalent linearized conductance. Thus:

$$P_{Si} = -\frac{a_{1s}^2}{2} G_{ei} = -\frac{a_{is}^2}{2} f_i(a_{1s}^2, a_{2s}^2, \cdots, a_{ns}^2). \qquad (25)$$

Also

$$P_{Di} = +\frac{a_{is}^2}{2} G_i. \qquad (26)$$

Now let

$$P_i = P_{Di} - P_{Si}.$$

In terms of these quantities the criteria (24) now become:

$$\text{(a)} \qquad \frac{\delta_{01}}{G_1} \frac{\partial P_1}{\partial(a_1^2)} + \frac{\delta_{02}}{G_2} \frac{\partial P_2}{\partial(a_2^2)} > 0,$$

$$\text{(b)} \qquad \frac{\partial P_1}{\partial(a_1^2)} \frac{\partial P_2}{\partial(a_2^2)} - \frac{\partial P_{S1}}{\partial(a_2^2)} \frac{\partial P_{S2}}{\partial(a_1^2)} > 0. \qquad (27)$$

The three criteria for the case of three degrees of freedom can be readily written down in similar manner, but to save space, only criterion (a) is given:

$$\frac{\delta_{01}}{G_1} \frac{\partial P_1}{\partial(a_1^2)} + \frac{\delta_{02}}{G_2} \frac{\partial P_2}{\partial(a_2^2)} + \frac{\delta_{03}}{G_3} \frac{\partial P_3}{\partial(a_3^2)} > 0. \qquad (28)$$

This is of the same form as (27)(a), but contains more terms. By dividing out the factor $(a_{11}+a_{22}+a_{33})$, (23)(b) is seen to be dimensionally a sum of double products of derivatives of power. It is thus of the same form as (27)(b), but also includes more terms. Criterion (23) (c) may be recognized as a new form of criterion, however, which is made up of terms which are triple products of derivatives of power. If we now recall the stability criterion for one degree of freedom, namely, $(\partial P_D/\partial_a) > (\partial P_S/\partial a)$ or equivalently, $(\partial P_D/\partial a^2) > (\partial P_S/\partial a^2)$, the trend of the stability criteria becomes evident. As the degrees of freedom are increased, so also are the stability criteria, the same number being required as there are degrees of freedom. Successive criteria which are added as required by additional degrees of freedom become successively more complex, while those criteria already established retain their basic form, but add more terms.

The first criterion is a relation between rates of change of power supplied and power dissipated. The other criteria apparently represent higher order energy balance and interchange relations among the various oscillations. The number of criteria appearing in a given case is equal to the number of degrees of freedom of the oscillatory circuit, provided the system is not in internal resonance, a condition in which the oscillations are synchronized and their frequencies are in ratios of small integers. In systems in which internal resonance occurs, additional criteria arise to represent the requirements for stability of synchronization. This has been studied by Fontana for two degrees of freedom with internal resonance.[6]

In conclusion it should be noted that the analysis presented here is of particular application to the stability of systems subjected to small perturbations. However, if the perturbation is large, the question of whether it will increase or will return to zero is unaffected unless the perturbation is so large as to carry the system beyond an adjacent limit cycle. In this case the perturbation cannot return to zero, and the response of the system is radically changed. Thus, in Fig. 2(c), a small perturbation of the amplitude about point 1 will always decay to zero. If the perturbation is large enough to carry the amplitude beyond point 2, however, the amplitude will continue to increase to the value corresponding to point 3. The limiting size of perturbation to which any amplitude may be subjected without crossing an adjacent limit cycle is, of course, known as soon as the equilibrium amplitudes have been calculated. If, however, the perturbation is large, but not large enough to cross an adjacent limit cycle, the system will behave qualitatively the same as it would for small perturbations. For this reason, the question of whether an equilibrium point is stable or unstable is best handled by assuming small perturbations and obtaining a system of linear differential equations. For the more general case in which perturbations may be large and precise, information is desired as to their manner of growth or decay, and the resulting differential equations describing them must be nonlinear. However, the nonlinearity involved is of the same type as that in the original problem of determining amplitudes of equilibrium, and the solution of this problem may be expected to be no more difficult than that of the amplitude problem. An interesting aspect of this problem which is currently being studied by several investigators is the determination of the optimum type of nonlinearity to be included in a system to obtain a particular desired behavior of the perturbed system.

ACKNOWLEDGMENT

The author wishes to acknowledge the encouragement and helpful guidance of Dr. Lloyd T. DeVore in the accomplishment of the work reported here.

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[6] R. E. Fontana, "Internal resonance in circuits containing nonlinear resistance," PROC. I.R.E., vol. 39, pp. 945–951; August, 1951.

The Constants in the Equation for Atmospheric Refractive Index at Radio Frequencies*

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***Summary*—Recent improvements in microwave techniques have resulted in precise measurements which indicate that the conventional constants $K_1=79°\mathrm{K/mb}$ and $K_2'=4{,}800°\mathrm{K}$ in the expression for the refractivity of air, $N=(n-1)10^6=[K_1/T](p+K_2'\,e/T)$ should be revised. Various laboratories appear to have arrived at this conclusion independently. In much of radio propagation work the absolute value of the refractive index of the atmosphere is of small moment. However, in some work it is important and it seems highly desirable to decide upon a particular set of constants.**

Through consideration of the various recent experiments this paper arrives at a relation

$$N=\frac{77.6}{T}\left(p+4{,}810\frac{e}{T}\right)$$

where

p = total pressure in millibars
e = partial pressure of water vapor in millibars
T = absolute temperature = °C + 273

This expression is considered to be good to 0.5 per cent in N for frequencies up to 30,000 mc and normally encountered ranges of temperatures, pressure and humidity.

RECENT IMPROVEMENTS in microwave techniques have resulted in measurements at the National Bureau of Standards,[1] the National Physical Laboratory[2] and elsewhere[3-5] which have indicated that the conventional constants in the expression for the refractive index of air at radio frequencies should be revised. Various laboratories appear to have arrived at this conclusion independently with the result that there are several different sets of constants in current use.[6-9] The sources of these recent changes, such as have been run to earth, have been found to be based on individual rather than collective results. Almost all of the proposed constants seem to represent a substantial improvement over the former values. The authors propose a set of constants derived from what is felt to be the most reliable of the recent microwave and optical measurements of the refractive index of dry air and from a recent survey of water vapor Debye constants. It is hoped these new constants will provide a common meeting ground for the laboratories desiring change rather than inject just another set of values into the field.

For an accuracy of 0.5% in N, the scaled-up refractivity of moist air [$N=(n-1)10^6$, where n is the refractive index] some simplifying assumptions may be made if the use of the relation is to be restricted to certain limits of the variables. The limits in this case restrict its use to temperature ranges of −50 to +40°C, total pressures of 200 to 1,100 mb, water vapor partial pressures of 0 to 30 mb and a frequency range of 0 to 30,000 mc. The constituents of dry air and even water vapor may be assumed to obey the ideal gas law.[10] The refractive index of water vapor, a polar molecule with an electric dipole, may be represented by a two-term Debye relation.[11] The permeability of air at radio frequencies due to oxygen may be taken[1] as $1+0.4\times10^{-6}$. Dispersion may be neglected. The Lorentz polarization term may be ignored.[12] Absolute zero for temperature may be taken as −273°C rather than −273.16°C.

There has been no proof of variation in the composition of the dry gases of the free atmosphere either with latitude or with height up to the ionosphere.[13] The water vapor content, of course, varies widely. As contributions to the total refractive index obey an additive rule, a three-term expression may be formulated[11] in which the first term expresses the sum of the distortions of electronic charges of the dry gas molecules under the influence of an applied electromagnetic field, the second term this distortion for water vapor, and the third term the effect of the orientation of the electric dipoles of water vapor under the influence of a field. Thus, using N for the scaled-up refractivity [$N=(n-1)10^6$]

$$N=K_1\frac{p_d}{T}+K_2\frac{e}{T}+K_3\frac{e}{T^2} \tag{1}$$

where:

n = the refractive index at radio frequencies
p_d = partial pressure of the dry gases
e = partial pressure of water vapor
T = absolute temperature

In radio work one is interested in propagation through the free atmosphere, therefore the composition of air should be taken to include an average amount of carbon dioxide. However, laboratory measurements usually are made on CO_2-free air due to variable concentrations of CO_2 in the laboratory. Hence the values of $\epsilon-1$ orig-

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* Decimal classification: R216.3. Original manuscript received by the Institute, October 10, 1952.

† Central Radio Propagation Laboratory, Boulder, Colorado.

[1] G. Birnbaum, S. J. Kryder and H. Lyons, "Microwave measurements of the dielectric properties of gases," *Jour. Appl. Phys.*, vol. 22, p. 95; 1951.

[2] L. Essen and K. D. Froome, "The refractive indices and dielectric constants of air and its principal constituents at 24,000 Mc/s," *Proc. Phys. Soc.* (London), B, vol. 64, p. 862; 1951.

[3] C. M. Crain, *Phys. Rev.*, vol. 74, p. 691; 1948.

[4] W. F. Gabriel, PROC. I.R.E., vol. 40, p. 940; August, 1952.

[5] W. E. Phillips, PROC. I.R.E., vol. 38, p. 786; 1950. See also comments by C. M. Crain, PROC. I.R.E., vol. 40, p. 164; February, 1952.

[6] C. I. Aslakson and C. O. Fickeissen, *Trans. Amer. Geoph. Un.*, vol. 31, p. 819; 1950.

[7] Research Interim Engineering Report No. 13, p. 23, Cornell University E. E.; October 23, 1951.

[8] L. Essen and K. D. Froome, *Nature*, vol. 167, p. 512; 1951.

[9] L. J. Anderson, NEL Report No. 279, page 3; 1952.

[10] H. Barrell and J. E. Sears, "On the refraction and dispersion of air for the visible spectrum," *Philos. Trans. A.*, vol. 238, p. 2; 1940.

[11] "Propagation of Short Radio Waves," Edited by D. E. Kerr, Radiation Laboratory Series, McGraw-Hill Book Co., Inc., N. Y., vol. 13, p. 189, 1951.

[12] *Ibid.*, p. 643.

[13] "Ionospheric radio propagation," National Bureau of Standards, Circ. 462, p. 35; June 1948.

inally published for CO_2-free air have been adjusted for 0.03% CO_2 content by raising them 0.02%. These values are also given on a real rather than an ideal gas basis. Three determinations shown in Table I are considered. The first shown, that of Barrell,[14] is an average of the constant term (n for $\lambda = \infty$) of the optical Cauchy dispersion equations for standard air used in three of the principal metrology laboratories of the world. Theoretical considerations indicate that the dielectric constant for dry air will be the same for optical and radio frequencies. Barrell's value is converted to dielectric constant from the relation $n = \sqrt{\mu\epsilon}$ with μ, the permeability taken as unity at optical frequencies. The second value, that of Birnbaum, Kryder and Lyons[1] was originally published on an ideal basis but has here been converted to the value pertinent to real gases. Last determination, (Essen and Froome[2]) has been adjusted to include CO_2. Uncertainties listed are standard errors.

TABLE I

DRY AIR REFRACTIVE INDEX AND DIELECTRIC CONSTANT AT 0°C AND 1 ATMOSPHERE

	Freq. of Measurement	Measured N	$(\epsilon - 1)10^6$	Year
1. Barrell[14]	Optical	$287.7_6 \pm 0.0_{65}$	$575.6 \pm 0.1_3$	1951
2. Birnbaum, Kryder and Lyons[1]	9,000 mc		$575.8 \pm 0.3_6$	1951
3 Essen and Froome[2]	24,000 mc	$288.2_1 \pm 0.1$	576.1 ± 0.2	1951
Mean Value of $(\epsilon - 1)10^6$			$575.7_5 \pm 0.1_0$	
Mean Value of N†		$288.0_4 \pm 0.0_5$		

† Derived from $n = \sqrt{\mu\epsilon}$ where $\mu - 1 = 0.4 \times 10^{-6}$ is taken for radio frequencies to account for the permeability.

The statistical mean value of dielectric constant is then converted to refractive index. The constant K_1 is evaluated from

$$N = K_1 \frac{p_d}{T} \qquad (2)$$

which stems from (1) when $e = 0$. Setting $N = 288.04$, $p = 1{,}013.25$ mb, $T = 273$°C and solving for K_1:

$$K_1 = 77.60_7 \pm 0.01_3 \frac{°\text{K}}{\text{mb}}. \qquad (3)$$

A recent survey of determinations of the dielectric constant of water vapor in the microwave region by Birnbaum and Chatterjee[15] is used to evaluate the water vapor constants K_2 and K^3 in (1). Assuming ideal gas behavior, which is permissible here as only low partial pressures are of interest, the constants may be evaluated from the Debye constants A and B (molar polarization $P = A + B/T$) determined by Birnbaum and Chatterjee. This results in:

$$K_2 = 71._6 \pm 8._5 \frac{°\text{K}}{\text{mb}} \qquad (4)$$

$$K_3 = (3.74_7 \pm 0.03_1)10^5 \frac{°\text{K}^2}{\text{mb}} \qquad (5)$$

where the uncertainties are again standard errors. Substituting the values of (3), (4) and (5) in (1) and reducing the values to three figures where significant

$$N \doteq 77.6 \frac{p_d}{T} + 72 \frac{e}{T} + 3.75 \times 10^5 \frac{e}{T^2}. \qquad (6)$$

Utilizing the total pressure $p = p_d + e$ one may write

$$N = 77.6 \frac{p}{T} - 6 \frac{e}{T} + 3.75 \times 10^5 \frac{e}{T^2}. \qquad (7)$$

TABLE II

TABLE OF CONSTANTS USED BY DIFFERENT AUTHORS

$$N = (n-1)10^6 = K_1 \frac{p}{T} + K_2 \frac{e}{T} + K_3 \frac{e}{T^2}$$

	Year	K_1	K_2	$K_3 \times 10^{-5}$
1. Schelleng, Burrows & Ferrell[16]	1933	79	675	1.35
2. Englund, Crawford & Mumford[17]	1935	79.1	68.3	3.81
3. Waynick[18]	1940	79	68.5	3.72
4. Smith-Rose & Stickland[19]	1943	79	68	3.77
5. Burrows & Attwood-NDRC[20]	1946	79	68	3.8
6. Met. Factors in Propagation[21]	1946	79	68	3.8
7. Kerr, *Rad. Labs.*, vol. 13[11]	1951	79	(79)	3.8
8. Essen & Froome[2]	1951	77.64	64.68	3.718
9. Smith & Weintraub	1952	77.6	72	3.75

For use in the limited temperature range −50°C to +40°C, negligible error is incurred through lumping the second and third terms in (7). This may be accomplished by dividing these terms by e/T and solving for the composite constant, K_4, in the relation:

$$\frac{3.75 \times 10^5}{T} - 6 = \frac{K_4}{T} \qquad (8)$$

which, for $T = 273$°K, results in:

$$K_4 = 3.73 \times 10^5. \qquad (9)$$

The two-term formula for moist air is now

$$N = 77.6 \frac{p}{T} + 3.73 \times 10^5 \frac{e}{T^2} \qquad (10)$$

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[14] H. Barrell, "The dispersion of air between 2500A and 6500A," *Jour. Opt. Soc. Amer.*, vol. 41, p. 295; 1951.

[15] G. Birnbaum and S. K. Chatterjee, "The dielectric constant of water vapor in the microwave region," *Jour. Appl. Phys.*, vol. 23, p. 220; February, 1952.

[16] J. C. Schelleng, C. R. Burrows, and E. B. Ferrell, "Ultra-short-wave propagation," PROC. I.R.E., vol. 21, p. 427; 1933.

[17] C. R. Englund, A. B. Crawford, and W. W. Mumford, "Further results of a study of ultra-short-wave transmission phenomena," *Bell Sys. Tech. Jour.*, vol. 14, p. 369; 1935.

[18] A. H. Waynick, "Experiments on the propagation of ultra-short radio waves," PROC. I.R.E., vol. 28, p. 468; 1940.

[19] R. L. Smith-Rose and A. C. Strickland, "A study of propagation over the ultra-short-wave radio link between Guernsey and England on wavelengths of 5 and 8 meters," *J.I.E.E.*, vol. 90, III, p. 12; 1943.

[20] "Radio Wave Propagation," Consolidated Summary Technical Report on the Committee on Propagation, NDRC. Edited by C. R. Burrows and S. S. Attwood, p. 219 (Academic Press Inc., New York, N. Y. 1949.)

[21] "Meteorological Factors in Radio Wave Propagation," *The Physical Society*, London: 1946. Foreword.

which may be written

$$N = \frac{77.6}{T}\left(p + 4.81 \times 10^3 \frac{e}{T}\right). \tag{11}$$

These last two relations are the ones proposed for general radio meteorological use.

Listed in Table II are some of the values of K_1, K_2 and K_3 that have been used by authors through the years. In examples 1 through 7 the constant K_1 was drawn from the Smithsonian Physical Tables (1933).

The water vapor constants K_2 and K_3 in examples 2 through 7 represent an average of the determinations of several different workers in each case. Essen and Froome in example 8 relied on their own excellent experimental work to evaluate K_1 and K_3. Their value of K_1 is seen from Table I to be in excellent agreement with the values of Barrell and of Birnbaum, Kryder and Lyons. However, their value for K_3 is low because of the fact that their measured value of the dielectric constant of water vapor is about 1% lower than those determined by other recent workers in the field. It is true that their measurements were performed in a region adjacent to the 1.35 cm water vapor absorption band, but this effect has been shown to be too small to account for the discrepancy. As Essen and Froome did not have refractive index determinations over the wide temperature range necessary to evaluate K_2 and K_3 independently, they utilized an optical value of refractive index ($K_3=0$) to determine K_2. Consequently, their value for K_2 is about 10% low at radio frequencies; however, the effect of this on the equation is negligible.

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Reduction of Forced Error in Closed-Loop Systems*

LEONARD H. KING†

Summary—**This paper illustrates a method for reducing the forced-dynamic error in servomechanisms by design based on error coefficients. After a brief review of the dependence of the forced dynamic error upon error coefficients, the relationship between error coefficients and the parameters of a servomechanism is derived. This relationship is then used to show how closed-loop systems can be modified to obtain favorable error coefficients that reduce the forced-dynamic error. The method has been tested by simulation, and photographs of simulator response show how the effect of additional integrations can be achieved by error-coefficient adjustment.**

I. Introduction

IN A PERFECT POSITIONAL SERVOMECHANISM, the output position would always be aligned with the input command position regardless of noise, disturbing torques, or variations in the input position. As it is impossible to build a perfect positional servomechanism, good design practice consists largely of trying to reduce the following components of error:

1. Noise error, due to the transmission of input and internal noise to the output.
2. Position error, caused by disturbing torques.
3. Transient error, due to initial conditions or input discontinuities. By definition, this error disappears in one response time after which the output remains within five per cent of the correct output for a step change in the input.[1]
4. Forced dynamic error, which exists after one response time and is due to variations in the input command position.

This paper illustrates a method for reducing the forced dynamic component of error by design based on error coefficients.[2-5] After a brief review of the dependence of the forced dynamic error upon error coefficients, the relationship between error coefficients and the parameters of a servomechanism is derived. This relationship is then used to show how closed-loop systems can be modified to obtain favorable error coefficients that reduce the forced dynamic error. The method has been tested by simulation, and photographs of simulator response show how the effect of additional integrations can be achieved by error coefficient adjustment. Specifically, it is shown (1) how a loop with no integrations can be made to have zero positional and velocity error and (2) how a loop with one integration can be made to have zero velocity and acceleration error.

II. Relationship between Forced-Dynamic Error and Error Coefficients

If the input to a servo system varies continuously with time, then there will be a forced dynamic error $E(t)$ between the desired input $q_{\text{in}}(t)$ and the actual system output $q_{\text{out}}(t)$. For finite derivatives of the input function, the forced dynamic error can generally be expanded in a series as

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[1] C. S. Draper, W. McKay, and S. Lees, "Instrument Engineering," vol. 1, McGraw-Hill Book Co., Inc., New York, N. Y.; 1952.

[2] H. M. James, N. B. Nichols, and R. S. Phillips, "Theory of Servomechanisms," vol. 25, Rad. Lab. Series, McGraw-Hill Book Co., Inc., New York, N. Y.; 1947.

[3] G. H. Floyd, Jr., and R. H. Eisengrein, "Design of Fire-Control Systems," Servomechanisms Laboratory, M.I.T., Cambridge, Mass. (Confidential), July, 1950.

[4] J. L. Bower, "A note on the error coefficients of a servomechanism," *Jour. Appl. Phys.*, vol. 21; July, 1950.

[5] H. Chestnut and R. W. Mayer, "Servomechanisms and Regulating System Design," vol. 1, John Wiley & Sons, Inc., New York, N. Y.; 1951.

$$E(t) = a_0 q_{\text{in}}(t) + a_1 \frac{d}{dt} q_{\text{in}}(t) + a_2 \frac{d^2}{dt^2} q_{\text{in}}(t) + \cdots + a_n \frac{d^n}{dt^n} q_{\text{in}}(t) + \cdots \tag{1}$$

where

$E(t)$ is the forced dynamic error.
$q_{\text{in}}(t)$ is the input function.
a_0 = zero-order (positional) error coefficient.
a_1 = first-derivative (velocity) error coefficient.
a_2 = second-derivative (acceleration) error coefficient.
a_n = nth-derivative error coefficient.

For example, if $q_{\text{in}}(t)$ can be represented as $1 + 0.1t + 0.01t^2$, the error as a function of time, after initial conditions have disappeared, can be evaluated completely as

$$E(t) = a_0(1 + 0.1t + 0.01t^2) + a_1(0.1 + 0.02t) + a_2(0.02).$$

It can be seen that the error coefficients lead directly to the evaluation of the resulting error for all values of time rather than only for time equal to infinity, as with error constants, which have been treated elsewhere in the literature.[6-8]

III. EVALUATION OF ERROR COEFFICIENTS

A perfect servo is defined as one having all error coefficients equal to zero. In practice, it is desired to make as many of the lower-order error coefficients zero as possible. This is equivalent to making the more familiar error constants K_p, K_v, K_a, etc., infinite.

The error coefficients can be calculated by considering the Laplace transform of the error between the input and the output of the system.

Let

$$E(s) = q_{\text{out}}(s) - q_{\text{in}}(s), \tag{2}$$

where $E(s)$ is the Laplace transform of the error $E(t)$ between the output and input quantities $q_{\text{out}}(t)$ and $q_{\text{in}}(t)$. Dividing both sides of (2) by $q_{\text{in}}(s)$ yields

$$\frac{E(s)}{q_{\text{in}}(s)} = \left[\frac{q_{\text{out}}(s)}{q_{\text{in}}(s)} - 1\right], \tag{3}$$

where $q_{\text{out}}(s)/q_{\text{in}}(s)$ is the closed-loop transfer function if zero initial conditions are assumed. Equation (3) can be expanded in a Maclaurin series for small values of s as

$$\frac{E}{q_{\text{in}}}(s) = a_0 + a_1 s + a_2 s^2 + \cdots + a_n s^n + \cdots. \tag{4}$$

From the similarity between (1) and (4), it can be seen that the a's are the so-called error coefficients.

It is possible to evaluate successive error coefficients from the following relationships derived from (4)

$$a_0 = \lim_{s\to 0}\left[\frac{E(s)}{q_{\text{in}}(s)}\right], \tag{5a}$$

$$a_1 = \lim_{s\to 0}\frac{1}{s}\left[\frac{E(s)}{q_{\text{in}}(s)} - a_0,\right] \tag{5b}$$

$$a_n = \lim_{s\to 0}\frac{1}{s^n}\left[\frac{E(s)}{q_{\text{in}}(s)} - \sum_{k=0}^{n-1} a_k s^k\right]. \tag{5c}$$

Instead of calculating error coefficients from the Maclaurin expansion of $[(q_{\text{out}}/q_{\text{in}})(s) - 1]$, it is easier and more instructive to calculate the values from $(q_{\text{out}}/q_{\text{in}})(s)$, which is the familiar closed-loop transfer function. From (3) and (4), it can be seen that

$$\frac{q_{\text{out}}}{q_{\text{in}}}(s) = 1 + a_0 + a_1 s + a_2 s^2 + \cdots + a_n s^n + \cdots. \tag{6}$$

Equation (6) is used to calculate the error coefficients. Let

$$a_0' = 1 + a_0 \tag{7}$$

and

$$\frac{q_{\text{out}}}{q_{\text{in}}}(s) = \frac{n_0 + n_1 s + n_2 s^2 + \cdots + n_n s^n}{1 + d_1 s + d_2 s^2 + \cdots + d_m s^m}. \tag{8}$$

Then, from (6), (7) and (8);

$$a_0' = \lim_{s\to 0}\left[\frac{q_{\text{out}}}{q_{\text{in}}}(s)\right] = n_0, \tag{9}$$

$$a_1 = \lim_{s\to 0}\frac{1}{s}\left[\frac{q_{\text{out}}}{q_{\text{in}}}(s) - a_0'\right] = n_1 - a_0' d_1, \tag{10}$$

$$a_2 = \lim_{s\to 0}\frac{1}{s^2}\left[\frac{q_{\text{out}}}{q_{\text{in}}}(s) - a_0' - a_1 s\right] = n_2 - a_0' d_2 - a_1 d_1, \tag{11}$$

$$a_n = \lim_{s\to 0}\frac{1}{s^n}\left[\frac{q_{\text{out}}}{q_{\text{in}}}(s) - a_0' - \sum_{k=1}^{n-1} a_k s^k\right] = n_n - a_0' d_n - \sum_{y=1}^{n-1} a_y d_{n-y}. \tag{12}$$

Equation (12) provides a simple method of computing higher-order error coefficients in terms of lower-order error coefficients. Moreover, (8) and (12) give an insight into how $(q_{\text{out}}/q_{\text{in}})(s)$ should be modified to provide the desired error coefficients. For example, with all the lower-order error coefficients up to a_n equal to zero, the coefficient a_n is made zero by making n_n equal d_n.

IV. METHODS OF OBTAINING ZERO ERROR COEFFICIENTS

To correlate the desired error coefficients with the parameters of $(q_{\text{out}}/q_{\text{in}})(s)$, two block diagrams are next examined in detail in order to see how error coefficients of order n and $n+1$ can be made equal to zero in single loops containing integrations of order n. Photographs of test results are shown.

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[6] H. M. James, et al., *loc cit.*
[7] H. Chestnut and R. W. Mayer, *loc cit.*
[8] G. S. Brown and D. P. Campbell, "Principles of Servomechanisms," John Wiley & Sons, Inc., New York, N. Y.; 1948.

A. Loop with No Integrations

First, it is shown how the the system of Fig. 1, which contains no integrations, can be modified to obtain zero positional and zero velocity error coefficients. With K_2 gain compensation inserted into the feedback loop of Fig. 1, $(q_{out}/q_{in})(s)$ can be written as follows

$$\frac{q_{out}}{q_{in}}(s) = \frac{s^2\left(\dfrac{K_1T_1T_3}{1+KK_2}\right) + s\left(\dfrac{KT_1+KT_3}{1+KK_2}\right) + \dfrac{K}{1+KK_2}}{s^2\left(\dfrac{T_2T_4+KK_2T_1T_3}{1+KK_2}\right) + s\left(\dfrac{T_2+T_4+KK_2T_1+KK_2T_3}{1+KK_2}\right)+1}. \tag{13}$$

Equations (7) and (9) show that for a_0 to be equal to zero, $(1+KK_2)$ must equal K. Therefore K_2 must equal $(1-(1/K))$, and with this value the system of Fig. 1 now can be classified as a positional servomechanism since it has zero positional error. The same system configuration with unity feedback, however, does not meet with the zero positional error requirement of a servo.

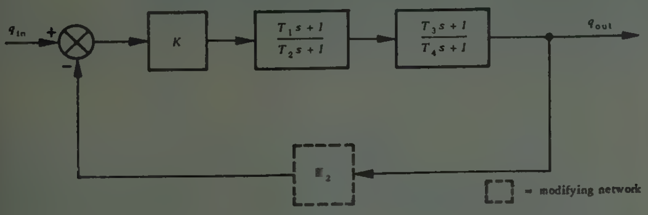


Fig. 1—General closed-loop system containing no integration.

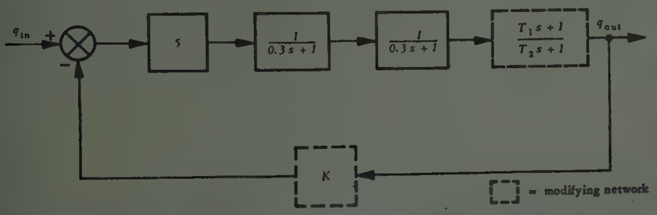


Fig. 2—Sample loop containing no integrations.

To determine a_1, (10) is now used

$$a_1 = \frac{KT_1+KT_3}{1+KK_2} - \frac{T_2+T_4+KK_2T_1+KK_2T_3}{1+KK_2}. \tag{14}$$

But

$$KK_2 = K - 1, \tag{15}$$

therefore

$$a_1 = \frac{K(T_1+T_3) - T_2 - T_4 - (K-1)(T_1+T_3)}{K} \tag{16}$$

$$= \frac{T_1+T_3-T_2-T_4}{K}. \tag{17}$$

By adding a network to the forward loop so that the sum of the time constants in the numerator equals the sum of the time constants in the denominator, the servo will now have a zero velocity error coefficient.

In order to illustrate the effectiveness of the proposed method of compensation, the block diagram of Fig. 2 was examined on the General-Purpose Simulator at the Instrumentation Laboratory, Massachusetts Institute of Technology, and photographs were taken of the transient response as seen on an oscilloscope. The pictures of Fig. 3 show that the forced-dynamic error of this specific system has been reduced by using eight-tenths feedback instead of unity feedback. Furthermore, the gain of the feedback loop K is not critical in the system. As long as K is between six-tenths and unity, the forced-dynamic error of the system would be less than that present with unity feedback.

It can also be seen from the pictures of Fig. 3 that the forced dynamic error for a ramp input has been reduced to zero by insertion of a passive network in

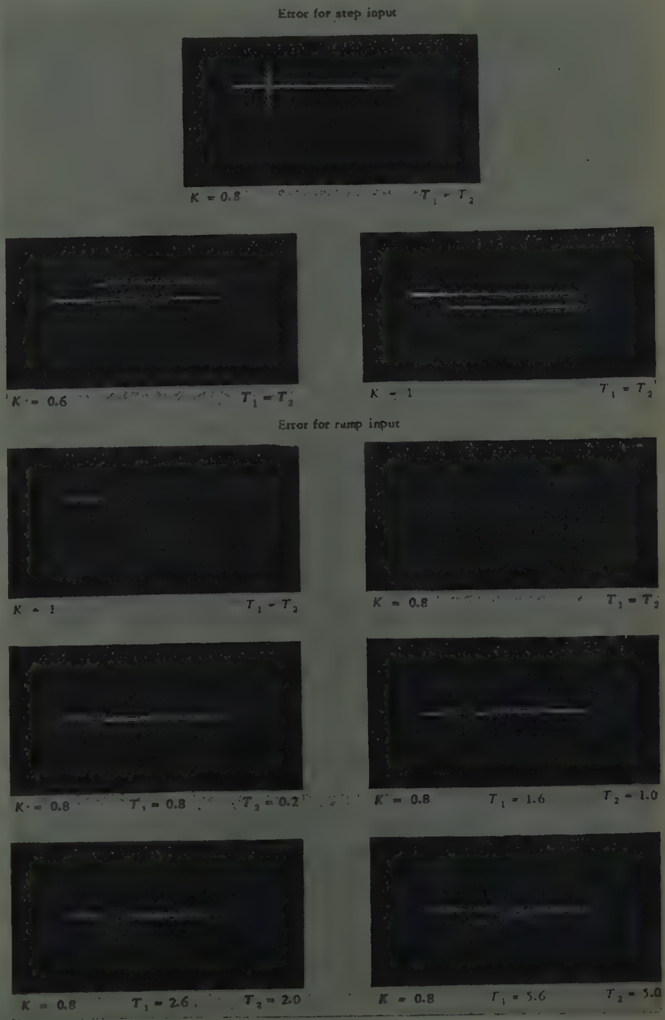
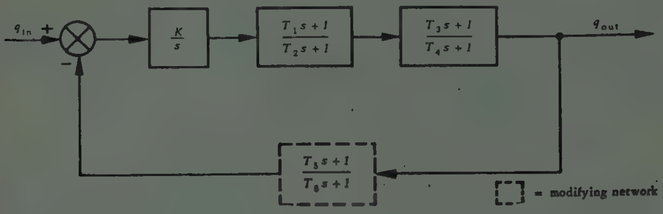


Fig. 3—Simulator results obtained for system of Fig. 2.

the forward path. As there is a wide choice in the parameters of the actual network used, the transient error responses are shown for several different networks.

B. Single Loop with One Integration

It will now be shown how the system of Fig. 4, which contains a single integration, can be modified to obtain zero velocity and acceleration coefficients.



q_{in}
+
−
$\frac{K}{s}$
$\frac{T_1 s+1}{T_2 s+1}$
$\frac{T_3 s+1}{T_4 s+1}$
q_{out}
$\frac{T_5 s+1}{T_6 s+1}$
= modifying network

Fig. 4—General closed-loop system containing one integration.

As a first modification, the network $(T_5s+1)/(T_6s+1)$ is placed in the feedback path. Then

$$\frac{q_{out}}{q_{in}}(s)=\frac{\frac{T_1s+1}{T_2s+1}\times\frac{T_3s+1}{T_4s+1}\times\frac{K}{s}}{1+\frac{T_1s+1}{T_2s+1}\times\frac{T_3s+1}{T_4s+1}\times\frac{T_5s+1}{T_6s+1}\times\frac{K}{s}}, \tag{18}$$

$$\frac{q_{out}}{q_{in}}(s)=\frac{n_0+n_1s+n_2s^2+n_3s^3}{1+d_1s+d_2s^2+d_3s^3+d_4s^4}, \tag{19}$$

where

$$\begin{aligned} n_0 &= 1\\ n_1 &= T_1+T_3+T_6\\ n_2 &= T_1T_3+T_1T_6+T_3T_6\\ d_1 &= T_1+T_3+T_5+\frac{1}{K}\\ d_2 &= T_1T_3+T_3T_5+T_1T_5+\frac{1}{K}(T_2+T_4+T_6). \end{aligned} \tag{20}$$

From these equations, used in conjunction with (12), it is seen that

$$a_0 = 0,$$

and

$$\begin{aligned} a_1 &= T_1+T_3+T_6-\left(T_1+T_3+T_5+\frac{1}{K}\right). \\ &= T_6-\left(T_5+\frac{1}{K}\right) \end{aligned} \tag{21}$$

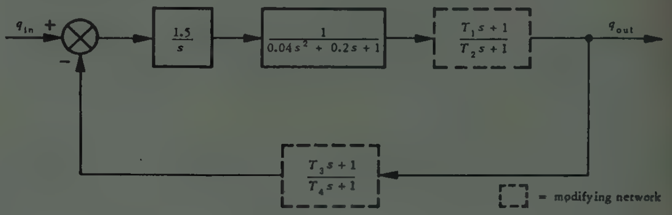
Therefore, a_1 is dependent only upon the parameters of the feedback path and the gain of the forward path. Specifically, in order for a_1 to be equal to zero, T_6 must equal $T_5+(1/K)$. With T_5 equal to zero and a_1 made equal to zero by the proper choice of T_6, a_2 is now given by

$$\begin{aligned} a_2 = {} & T_1T_3+T_1T_6+T_3T_6-T_1T_3\\ & -\frac{1}{K}(T_2+T_4+T_6). \end{aligned} \tag{22}$$

Because $T_6=1/K$, then

$$a_2=\frac{T_1}{K}+\frac{T_3}{K}-\frac{T_2}{K}-\frac{T_4}{K}-\frac{1}{K^2}. \tag{23}$$

With a_1 equal to zero, a_2 can, therefore, be made equal to zero by ensuring that in the forward part of the loop the sum of the first-order time constants in the numerator is equal to the sum of the first-order time constants in the denominator plus $1/K$.



q_{in}
+
−
$\frac{1.5}{s}$
$\frac{1}{0.04s^2+0.2s+1}$
$\frac{T_1 s+1}{T_2 s+1}$
q_{out}
$\frac{T_3 s+1}{T_4 s+1}$
= modifying network

Fig. 5—Sample loop containing one integration.

To illustrate further the results obtainable by the use of error coefficients, the block diagram of Fig. 5 was also examined on the General Purpose Simulator. The results, summarized in Fig. 6, show that it is possible to obtain zero velocity and zero acceleration error coefficients without appreciably changing the stability or response time of the system.

C. Generalized Technique

The effect of a servo in the feedback loop, such as in Fig. 7, is considered next. For a constant angular velocity input q_{in}, the following steady-state equations can be written

$$q_1 = q_{in}-\frac{\dot{q}_{in}}{K} \tag{24}$$

$$q_1 = q_{out}-\frac{\dot{q}_{out}}{K_A} \tag{25}$$

$$\dot{q}_{in}=\dot{q}_{out}. \tag{26}$$

For q_{out} to be equal to q_{in}, K_A must therefore equal K. However, from the viewpoint of the over-all loop, the servo in the feedback path is equal to the passive network $1/[1+(1/K)s]$, which is identical with the modifying network that was calculated for the servo in Fig. 4. Consequently, the modifying passive network acts as though it were a servo having the proper velocity constant. In fact, any passive network which has a transfer function equal to the closed-loop transfer function of a servo having the correct velocity constant can be used in the feedback path of Fig. 4 to obtain a zero velocity error coefficient. In general, this analog method can be used for selecting the network necessary to make a_n equal to zero in a loop containing n integrations where n is 1 or larger.

D. Alternate Compensation Techniques

Previously in this paper, emphasis has been placed on using a network in the feedback path in order to make the a_n error coefficient equal to zero in an n integration loop. In many cases, it may be impossible or inconven-

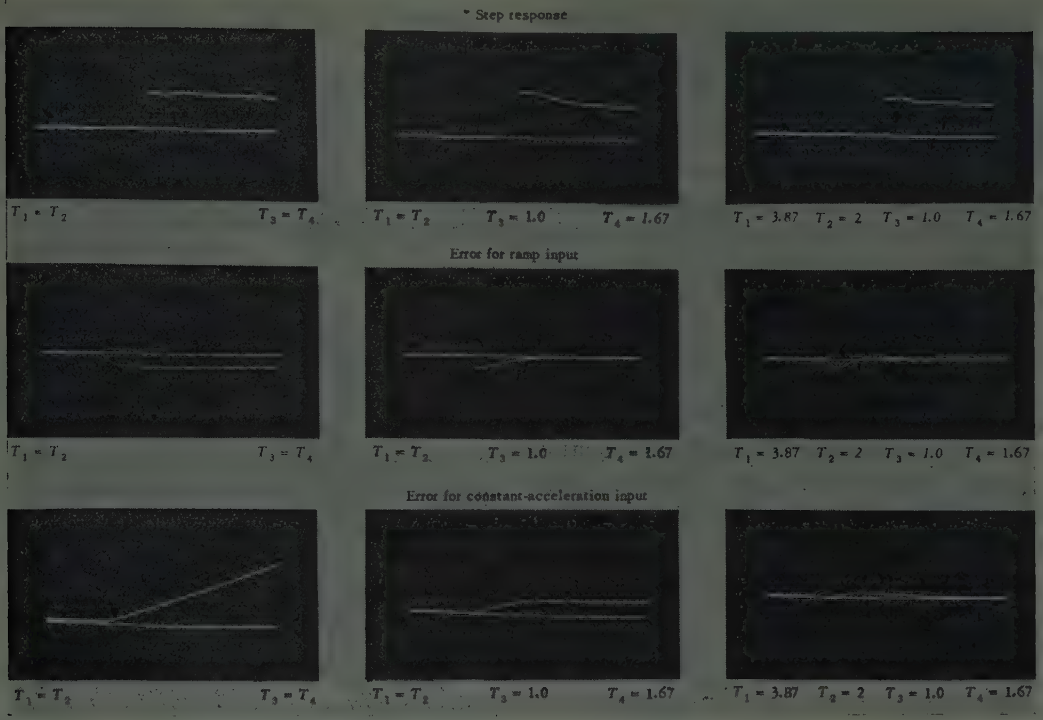


Fig. 6—Simulator results obtained for system of Fig. 5.

ient to put a network in the feedback path, due to the nature of the system components. However, even with a unity feedback path, it may still be possible to have the desired error coefficient. Fig. 8 shows two possible ways in which this may be done. In multiloop systems more ways are possible and the equivalent method of compensation will depend, to a large extent, upon the ingenuity of the designer.

E. Use of Additional Integrations

In a single-loop system with unity feedback, higher-order error coefficients can be made equal to zero by the insertion of additional integrations in the forward path. Thus, the compensation techniques developed in this paper are mathematically equivalent to the use of extra integrations in the forward path of a unity feedback system. For example, the system of Fig. 5, which contains only one integration and two passive compensating networks, is mathematically equivalent to a system with unity feedback and three integrations in the forward path. In general, each additional passive compensation network in a single-loop system can be replaced by an active integrator. However, in multiloop systems, it may be possible to have one network take the place of several integrators.

V. Effect of Zero Error Coefficient Modifications upon System Response

A. Stability and Bandwidth

It has been shown in the preceding section that the passive network used in the feedback loop is equivalent to a servo. If the bandwidth of the equivalent servo is sufficiently larger than the bandwidth of the original

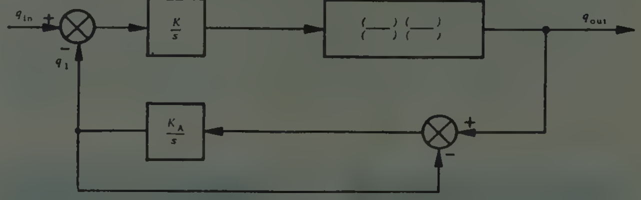

q_{in}
+
−
q_1
$\frac{K}{s}$
$\frac{(\quad)}{(\quad)}\frac{(\quad)}{(\quad)}$
q_{out}
$\frac{K_A}{s}$
+
−

Fig. 7—Example of system containing a servo in the feedback path

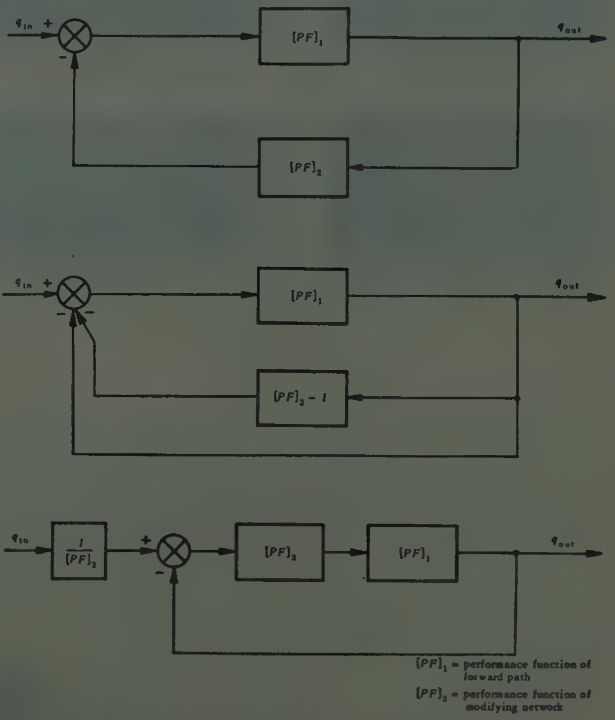

q_{in}
+
−
$[PF]_1$
q_{out}
$[PF]_2$
q_{in}
+
−
−
$[PF]_1$
q_{out}
$[PF]_2 - 1$
q_{in}
$\frac{1}{[PF]_2}$
+
−
$[PF]_2$
$[PF]_1$
q_{out}
$[PF]_1$ = performance function of forward path
$[PF]_2$ = performance function of modifying network

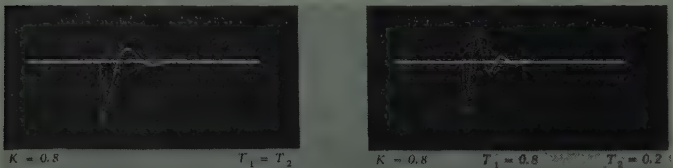
Fig. 8—Equivalent methods of compensation.

loop, then the over-all loop will see unity feedback inside its bandwidth. Since unity feedback does not change the phase lag at the resonant frequency, the passive network will not appreciably affect the stability or bandwidth of the over-all loop. Therefore, it is generally possible to design the passive network used in the feedback path so that it does not adversely affect the stability or bandwidth of the over-all loop. Moreover, if the original loop has marginal stability as a result of trying to achieve a large error constant, then both the stability and error constants of the loop can be increased by reducing the forward loop gain and using a network in the feedback loop.

The modifying network used in the forward loop has the function of making the a_{n+1} error coefficient equal to zero in an n integration loop. In a zero integration loop, this network should be such that the sum of the numerator time constants is equal to the sum of the denominator time constants. Since only the over-all sum is important, the networks shown below can be considered equivalent in the steady state.

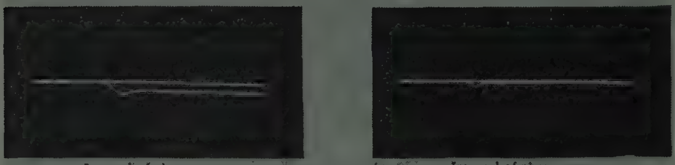
$$\frac{5s+1}{7s+1} \quad \text{or} \quad \frac{1s+1}{3s+1} \quad \text{or} \quad \frac{3s+1}{6s+1} \times \frac{2s+1}{1s+1}.$$

Consequently, there is a wide latitude in the choice of the modifying network, and it may be possible to pick the forward loop network so that it does not affect the stability or bandwidth. Likewise, in higher-order systems there is a wide latitude in the choice of the modifying network.



Error for step input

Error for step input



Integral of above error

Integral of above error

Fig. 9—Effect of modifications on transient response of system of Fig. 2.

B. *Reshaping of Transient Response*

It is known that the set of linear differential operators with constant coefficients is commutative.[9] Consequently, since a ramp function is the integral of a step function, the error response to a ramp input of a linear system with constant coefficients can be found by integrating the error response of the system for a step input. Similarly, the error response for a constant acceleration input is the double integral of the error response for a step input. Making higher-order error coefficients equal to zero, therefore, means that the error response to a step input is modified so that higher integrals of this response become equal to zero. This is illustrated in the pictures of Fig. 9.

VI. CONCLUSION

A method has been presented for reducing the forced dynamic error in a closed-loop system. In general, there is a wide latitude in the specific parameters of the networks used to accomplish this effect, and it is possible to design a network that does not appreciably change the stability or response time of the system.

Furthermore, even though it is a matching method of compensation that assumes the system parameters do not change, significant reduction of forced dynamic error can often be obtained without the network parameters having to be set precisely. For example, if a network $1/(T_7s+1)$ is used instead of the theoretically correct network $1/(T_6s+1)$ in the feedback path of Fig. 4, then the forced dynamic error for a ramp input will be less than would be present with unity feedback as long as $0<T_7<2T_6$. The choice of the network can be facilitated by the following general rule:

> In a single loop containing n integrations, the error coefficient a_n is independent of the dynamic lags of the forward path. It does, however, depend upon the gain of the forward path and the parameters of the feedback path. Specifically, in a loop containing no integrations, a zero positional error coefficient can be obtained by using a feedback path gain of one minus the reciprocal of the gain of the forward path. In a loop containing one integration, a zero velocity error coefficient can be obtained by using the feedback network $(T_5s+1)/(T_6s+1)$, where T_6 equals T_5 plus the reciprocal of the gain of the forward path.

Although this paper is restricted to the consideration of positional servomechanisms, the techniques developed here are applicable to a larger class of systems. If it is desired that the output q_{out} be some function of the input $f(q_{\text{in}})$, then the system error may be defined as $q_{\text{out}}-f(q_{\text{in}})$. With this definition of error, a new set of error coefficients may be formulated in terms of the system parameters. The system parameters may then be modified intelligently to reduce the forced dynamic error. Finally, since the compensation used is contingent upon modifying the feedback loop, it has been shown conclusively that servo performance can be improved by using other than unity feedback.

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[9] F. B. Hildebrand, "Advanced Calculus for Engineers," Prentice-Hall, Inc., New York, N. Y.; 1949.

Theory of AFC Synchronization*

WOLF J. GRUEN†, MEMBER, IRE

Summary—**The general solution for the important design parameters of an automatic frequency and phase-control system is presented. These parameters include the transient response, frequency response and noise bandwidth of the system, as well as the hold-in range and pull-in range of synchronization.**

I. INTRODUCTION

AUTOMATIC FREQUENCY and phase-control systems have been used for a number of years for the horizontal-sweep synchronization in television receivers, and more recently have found application for the synchronization of the color subcarrier in the proposed NTSC color-television system. A block diagram of a general AFC system is shown in Fig. 1.

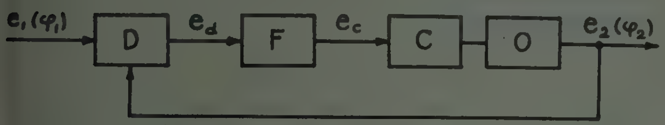

Fig. 1—Block diagram of A.F.C. loop.

The phase of the transmitted synchronizing signal e_1 is compared to the phase of a local oscillator signal e_2 in a phase discriminator D. The resulting discriminator output voltage is proportional to the phase difference of the two signals, and is fed through a control network F to a frequency-control stage C. This stage controls the frequency and phase of a local oscillator O in accordance with the synchronizing information, thereby keeping the two signals in perfect synchronism. Although in practice the transmitted reference signal is often pulsed and the oscillator comparison voltage non-sinusoidal, the analysis is carried out for sinusoidal signal voltages. The theory, however, can be extended for a particular problem by writing the applied voltages in terms of a Fourier series instead of the simple sine function. An AFC system is essentially a servomechanism, and the notation that will be used is the one followed by many workers in this field. An attempt will be made to present the response characteristics in dimensionless form in order to obtain a universal plot of the response curves.

II. DERIVATION OF THE BASIC EQUATION

If it is assumed that the discriminator is a balanced phase detector composed of peak-detecting diodes, the discriminator-output voltage can be derived from the vector diagram in Fig. 2. For sinusoidal variation with time, the synchronizing signal e_1 and the reference signal e_2 can be written

$$e_1 = E_1 \cos \phi_1 \tag{1}$$

and

$$e_2 = E_2 \sin \phi_2. \tag{2}$$

ϕ_1 and ϕ_2 are functions of time and, for reasons of simplicity in the later development, it is arbitrarily assumed that ϕ_1 and ϕ_2 are in quadrature when the system is perfectly synchronized, that is when $\phi_1 = \phi_2$.

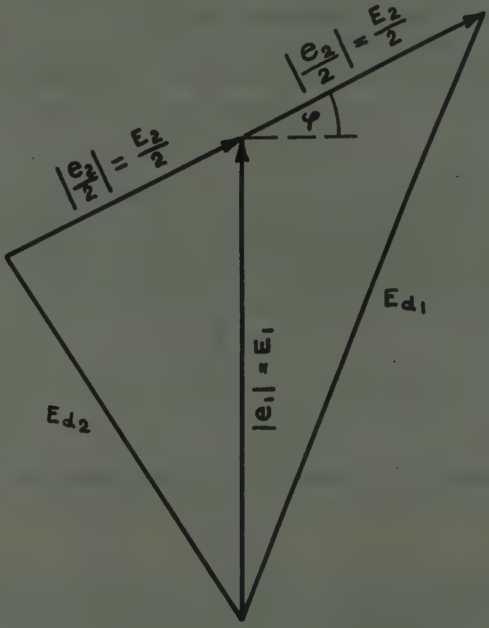
$\left|\frac{e_2}{2}\right| = \frac{E_2}{2}$
φ
$\left|\frac{e_2}{2}\right| = \frac{E_2}{2}$
E_{d_1}
$|e_1| = E_1$
E_{d_2}

Fig. 2—Discriminator vector diagram.

While one of the discriminator diodes is fed with the sum of e_1 and $e_2/2$, the other is fed with the difference of these two vectors as shown in Fig. 2. The resulting rectified voltages E_{d1} and E_{d2} can be established by simple trigonometric relations. Defining a difference phase

$$\phi \equiv \phi_1 - \phi_2, \tag{3}$$

one obtains

$$E_{d1}^2 = E_1^2 + \frac{E_2^2}{4} + E_1E_2 \sin \phi \tag{4}$$

and

$$E_{d2}^2 = E_1^2 + \frac{E_2^2}{4} - E_1E_2 \sin \phi. \tag{5}$$

The discriminator output voltage e_d is equal to the dif-

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* Decimal classification: R583.5. Original manuscript received by the Institute, August 21, 1952; revised manuscript received February 25, 1953.

† General Electric Co., Syracuse, N. Y.

ference of the two rectified voltages, so that

$$e_d = E_{d1} - E_{d2} = \frac{2E_1E_2}{E_{d1} + E_{d2}} \sin \phi. \tag{6}$$

If the amplitude E_1 of the synchronizing signal is larger than the amplitude E_2 of the reference signal, one obtains

$$E_{d1} + E_{d2} \cong 2E_1. \tag{7}$$

The discriminator output voltage then becomes

$$e_d = E_2 \sin \phi \tag{8}$$

and is independent of the amplitude E_1 of the synchronizing signal. As ϕ_1 and ϕ_2 are time-varying parameters, it should be kept in mind that the discriminator time constant ought to be shorter than the reciprocal of the highest difference frequency $d\phi/dt$, which is of importance for the operation of the system.

Denoting the transfer function of the control network F as $F(p)$, the oscillator control voltage becomes

$$e_c = F(p)E_2 \sin \phi. \tag{9}$$

Assuming furthermore that the oscillator has a linear-control characteristic of a slope S, and that the free-running oscillator frequency is ω_0, the actual oscillator frequency in operational notation becomes

$$p\phi_2 = \omega_0 + Se_c. \tag{10}$$

Substituting (3) and (9) into (10) then gives

$$p\phi + SE_2F(p) \sin \phi = p\phi_1 - \omega_0. \tag{11}$$

The product SE_2 repeats itself throughout this paper and shall be defined as the gain constant

$$K \equiv SE_2. \tag{12}$$

K represents the maximum frequency shift at the output of the system per radian phase shift at the input. It has the dimension of radians/second.

Equation (11) can be simplified further by measuring the phase angles in a coordinate system which moves at the free-running speed ω_0 of the local oscillator. One obtains

$$\boxed{p\phi + F(p)K \sin \phi = p\phi_1}. \tag{13}$$

This equation represents the general differential equation of the AFC feedback loop. $p\phi$ is the instantaneous-difference frequency between the synchronizing signal and the controlled-oscillator signal and $p\phi_1$ is the instantaneous-difference frequency between the synchronizing signal and the free-running oscillator signal.

Equation (13) shows that all AFC systems with identical gain constants K and unity d.c. gain through the control network have the same steady-state solution, provided that the difference frequency $p\phi_1$ is constant. If this difference frequency is defined as

$$\Delta\omega \equiv p\phi_1 = \omega_1 - \omega_0, \tag{14}$$

the steady-state solution is

$$\sin \phi = \frac{\Delta\omega}{K} \cdot \tag{15}$$

This means the system has a steady-state phase error which is proportional to the initial detuning $\Delta\omega$ and inversely proportional to the gain constant K. Since the maximum value of $\sin \phi$ in (15) is ± 1, the system will hold synchronism over a frequency range

$$\left| \Delta\omega_{\text{Hold-in}} \right| \leqq K. \tag{16}$$

Equations (15) and (16) thus define the static performance limit of the system.

III. Linear Analysis

An AFC system, once it is synchronized, behaves like a low-pass filter. To study its performance it is permissible, for practical signal-to-noise ratios, to substitute the angle for the sine function in (13). Then, with the definition of (3), one obtains

$$p\phi_2 + KF(p)\phi_2 = KF(p)\phi_1. \tag{17}$$

This equation relates the output phase ϕ_2 of the synchronized system to the input phase ϕ_1. It permits an evaluation of the behavior of the system to small disturbances of the input phase, if the transfer function $F(p)$ of the control network is specified.

a. $F(p) = 1$

This is the simplest possible AFC system, and represents a direct connection between the discriminator output and the oscillator control stage. Equation (17) then becomes

$$p\phi_2 + K\phi_2 = K\phi_1. \tag{18}$$

If the initial detuning is zero, the transient response of the system to a sudden step of input phase $|\phi_1|$ is

$$\frac{\phi_2}{|\phi_1|}(t) = 1 - e^{-Kt}. \tag{19}$$

Likewise, the frequency response of the system to a sine-wave modulation of the input phase is

$$\frac{\phi_2}{\phi_1}(j\omega) = \frac{1}{1 + j\dfrac{\omega}{K}} \cdot \tag{20}$$

The simple AFC system thus behaves like an RC-filter and has a cut-off frequency of

$$\omega_c = K \text{ [radians/sec]}. \tag{21}$$

George[1] has shown that the m.s. phase error of the system under the influence of random interference is proportional to the noise bandwidth, which is defined as

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[1] T. S. George, "Synchronizing systems for dot interlaced color TV," Proc. I.R.E.; February, 1951.

$$B = \int_{-\infty}^{+\infty} \left| \frac{\phi_2}{\phi_1}(j\omega) \right|^2 d\omega. \qquad (22)$$

The integration has to be carried out from $-\infty$ to $+\infty$ since the noise components on both sides of the carrier are demodulated. Inserting (20) into (22) then yields

$$B = \pi K \text{ [radians/sec]}. \qquad (23)$$

It was shown in (15) that for small steady-state phase errors due to average frequency drift, the gain constant K has to be made as large as possible, while now for good noise immunity, i.e., narrow bandwidth, the gain constant has to be made as small as possible. A proper compromise of gain then must be found to insure adequate performance of the system for all requirements. This difficulty, however, can be overcome by the use of a more elaborate control network.

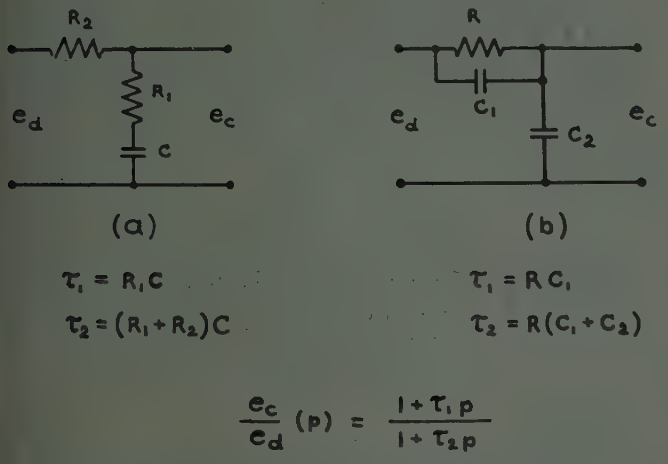

Fig. 3—Proportional plus integral control networks.

b. $F(p) = \dfrac{1 + \tau_1 p}{1 + \tau_2 p}$

Networks of this type are called proportional-plus integral-control networks[2] and typical network configurations are shown in Fig. 3. Inserting the above transfer function into (17) yields

$$p^2\phi_2 + \left(\frac{1}{\tau_2} + K\frac{\tau_1}{\tau_2}\right) p\phi_2 + \frac{K}{\tau_2}\phi_2 = K\frac{\tau_1}{\tau_2}\, p\phi_1 + \frac{K}{\tau_2}\phi_1. \qquad (24)$$

ϕ_1 and ϕ_2 are again relative phase angles, measured in a coordinate system which moves at the free-running speed of the local oscillator. To integrate (24), it is convenient to introduce the following parameters

$$\omega_n^2 \equiv \frac{K}{\tau_2} \qquad (25)$$

and

$$2\zeta\omega_n \equiv \frac{1}{\tau_2} + K\frac{\tau_1}{\tau_2}. \qquad (26)$$

ω_n is the resonance frequency of the system in the absence of any damping, and ζ is the ratio of actual-to-critical damping. In terms of the new parameters the time constants of the control network are

$$\tau_1 = \frac{2\zeta}{\omega_n} - \frac{1}{K} \qquad (27)$$

and

$$\tau_2 = \frac{K}{\omega_n^2}. \qquad (28)$$

With these definitions (24) becomes

$$p^2\phi_2 + 2\zeta\omega_n p\phi_2 + \omega_n^2\phi_2 = \left(2\zeta\omega_n - \frac{\omega_n^2}{K}\right) p\phi_1 + \omega_n^2\phi_1. \qquad (29)$$

The transient response of the system to a sudden step of input phase $|\phi_1|$ is found by integration of (29) and the initial condition for the oscillator frequency is obtained from (10). The transient response then is

$$\frac{\phi_2}{|\phi_1|}(t) = 1 - e^{-\zeta\omega_n t}\left[\cos\sqrt{1-\zeta^2}\,\omega_n t - \frac{\zeta - \dfrac{\omega_n}{K}}{\sqrt{1-\zeta^2}} \sin\sqrt{1-\zeta^2}\,\omega_n t\right]. \qquad (30)$$

For $\zeta < 1$ the system is underdamped (oscillatory), for $\zeta = 1$ critically damped and for $\zeta > 1$ overdamped (non-oscillatory). In order to avoid sluggishness of the system, a rule of thumb may be followed making. $4 < \zeta < 1^2$. The transient response of (30) can be plotted in dimensionless form if certain specifications are made for the ratio ω_n/K. As the time constant τ_1 of the control network must be positive or can at most be equal to zero, the maximum value for ω_n/K is found from (27), yielding

$$\left.\frac{\omega_n}{K}\right|_{\max} = 2\zeta. \qquad (31)$$

In this case the control network is reduced to a single time constant network ($\tau_1 = 0$). On the other hand, if for a fixed value of ω_n the gain of the system is increased towards infinity, the minimum value for ω_n/K becomes

$$\left.\frac{\omega_n}{K}\right|_{\min} = 0. \qquad (32)$$

Fig. 4 shows the transient response of the system for these two limits and for a damping ratio of $\zeta = 0.5$.

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[2] G. S. Brown and D. P. Campbell, "Principles of Servomechanisms," John Wiley & Sons Publishing Co., New York, N. Y.; 1948.

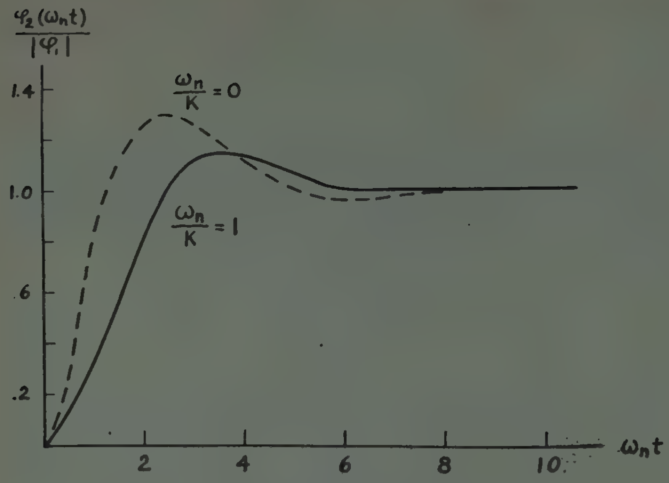

Fig. 4—Transient response for $\zeta=0.5$.

The frequency response of the system is readily found from (24) and one obtains

$$\frac{\phi_2}{\phi_1}(j\omega) = \frac{1 + j2\zeta\frac{\omega}{\omega_n}\left(1 - \frac{\omega_n}{2\zeta K}\right)}{1 + j2\zeta\frac{\omega}{\omega_n} - \left(\frac{\omega}{\omega_n}\right)^2}. \tag{33}$$

Its magnitude is plotted in Fig. 5 for the two limit values of ω_n/K and for a damping ratio $\zeta=0.5$. The curves show that the cut-off frequency of the system, for $\zeta=0.5$, is approximately

$$\omega_c \cong \omega_n \text{ [radians/sec.]}. \tag{34}$$

If ϕ_1 and ϕ_2 in (33) are assumed to be the input and output voltage of a four-terminal low-pass filter, the frequency response leads to the equivalent circuit of Fig. 6.

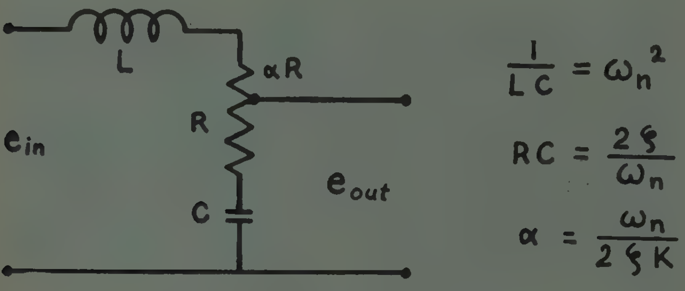
L
αR
R
C
e_{in}
e_{out}
$\frac{1}{LC}=\omega_n^2$
$RC=\frac{2\zeta}{\omega_n}$
$\alpha=\frac{\omega_n}{2\zeta K}$

Fig. 6—Equivalent low-pass filter.

The noise bandwidth of the system is established by inserting (33) into (22) and one obtains

$$B=\omega_n\int_{-\infty}^{+\infty}\frac{1+4\zeta^2\left(\frac{\omega}{\omega_n}\right)^2\left[1-\frac{\omega_n}{2\zeta K}\right]^2}{1-(2-4\zeta^2)\left(\frac{\omega}{\omega_n}\right)^2-\left(\frac{\omega}{\omega_n}\right)^4}\,d\left(\frac{\omega}{\omega_n}\right). \tag{35}$$

The integration, which can be carried out by partial fractions with the help of tables, yields

$$B = \frac{4\zeta^2 - 4\zeta\frac{\omega_n}{K} + \left(\frac{\omega_n}{K}\right)^2 + 1}{2\zeta}\pi\omega_n. \tag{36}$$

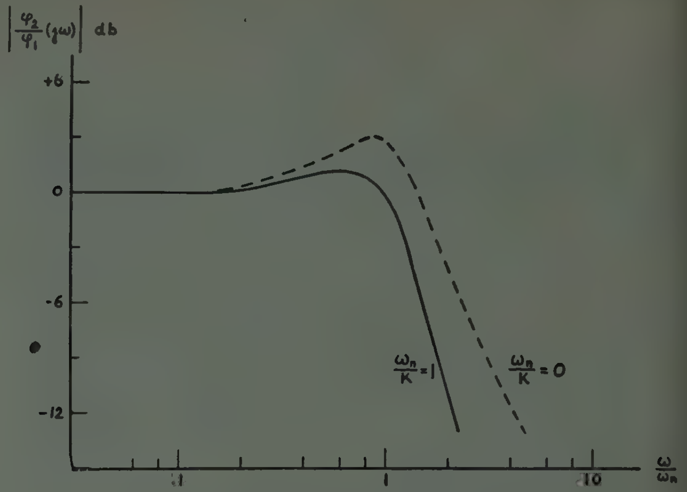

Fig. 5—Frequency response for $\zeta=0.5$.

For small values of ω_n/K, it is readily established that this expression has a minimum when $\zeta=0.5$. Hence, the noise bandwidths for the limit values of ω_n/K and $\zeta=0.5$ become

$$B\,\Big|_{(\omega_n/K)\to 1} = \pi\omega_n = \pi K \text{ [radians/sec]} \tag{37}$$

and

$$B\,\Big|_{(\omega_n/K)\to 0} = 2\pi\omega_n \text{ [radians/sec]}. \tag{38}$$

The above derivations, as well as the response curves of Figs. 4 and 5, show that the bandwidth and the gain constant of the system can be adjusted independently if a double time-constant control network is employed.

c. Example

The theory is best illustrated by means of an example. Suppose an AFC system is to be designed, having a steady state phase error of not more than 3° and a noise bandwidth of 1,000 cps. The local oscillator drift shall be assumed with 1,500 cps.

The required gain constant is obtained from (15), yielding

$$K = \frac{\Delta\omega}{\sin\phi} = \frac{2\pi\cdot 1{,}500}{\sin 3°} = 180{,}000 \text{ radians/sec.}$$

Since K is large in comparison to the required bandwidth, the resonance frequency of the system is established from (38).

$$\omega_n = \frac{B}{2\pi} = \frac{2\pi\cdot 1{,}000}{2\pi} = 1{,}000 \text{ radians/sec.}$$

The two time constants of the control network, assuming a damping ratio of 0.5, are determined from (27) and (28) respectively

$$\tau_1 = \frac{2\zeta}{\omega_n} - \frac{1}{K} = \frac{1}{1{,}000} - \frac{1}{180{,}000} \cong 10^{-3} \text{ sec,}$$

and

$$\tau_2 = \frac{K}{\omega_n^{\,2}} = \frac{180{,}000}{1{,}000^2} = 0.18 \text{ sec.}$$

These values K, τ_1, and τ_2 completely define the AFC system. A proper choice of gain distribution and control-network impedance still has to be made to fit a particular design. For example, if the peak amplitude of the sinusoidal oscillator reference voltage is $E_2=6$ volts, the sensitivity of the oscillator control stage must be $S=30{,}000$ radians/sec/volt to provide the necessary gain constant of 180,000 radians/sec. Furthermore, if the capacitor C for the control network of Fig. 3(a) is assumed to be 0.22 *uf*, the resistors R_1 and R_2 become 4.7 kΩ and 820 kΩ respectively, to yield the desired time constants.

IV. Non-Linear Analysis

While it was permissible to assume small phase angles for the study of the synchronized system, thereby linearizing the differential (13), this simplification cannot be made for the evaluation of the pull-in performance of the system. The pull-in range of synchronization is defined as the range of difference frequencies, $p\phi_1$, between the input signal and the free-running oscillator signal, over which the system can reach synchronism. Since the difference phase ϕ can vary over many radians during pull-in, it is necessary to integrate the nonlinear equation to establish the limit of synchronization.

Assuming that the frequency of the input signal is constant as defined by (14), (13) can be written

$$p\phi + F(p)K\sin\phi = \Delta\omega. \qquad (39)$$

Mathematically then, the pull-in range of synchronization is the maximum value of $\Delta\omega$ for which, irrespective of the initial condition of the system, the phase difference ϕ reaches a steady state value. To solve (39), the transfer function of the control network again must be defined.

a. $F(p) = 1$

The pull-in performance for this case has been treated in detail by Labin.[3] With $F(p)=1$ (39) can be integrated by separation of the variables and it is readily found that the system synchronizes for all values of $|\Delta\omega| < K$. The condition for pull-in then is

$$|\Delta\omega|_{\text{Pull-in}} < K. \qquad (40)$$

Large pull-in range and narrow-noise bandwidth thus are incompatible requirements for this system.

b. $F(p) = \dfrac{1+\tau_1 p}{1+\tau_2 p}$

Inserting this transfer function into (39) and carrying out the differentiation yields

$$\frac{d^2\phi}{dt^2} + \left[\frac{1}{\tau_2} + K\frac{\tau_1}{\tau_2}\cos\phi\right]\frac{d\phi}{dt} + \frac{K}{\tau_2}\sin\phi = \frac{\Delta\omega}{\tau_2}. \qquad (41)$$

This equation can be simplified by inserting the coefficients defined in (25) and (26), and by dividing the resulting equation by ω_n^2. This leads to the dimensionless equation.

$$\frac{d^2\phi}{\omega_n^2 dt^2} + \left[\frac{\omega_n}{K} + \left(2\zeta - \frac{\omega_n}{K}\right)\cos\phi\right]\frac{d\phi}{\omega_n dt} + \sin\phi = \frac{\Delta\omega}{K}. \qquad (42)$$

A further simplification is possible by defining a dimensionless difference frequency

$$y \equiv \frac{d\phi}{\omega_n dt} \qquad (43)$$

and one obtains a first order differential equation from which the dimensionless time $\omega_n t$ has been eliminated. It follows

$$\frac{dy}{d\phi} = \frac{\dfrac{\Delta\omega}{K} - \sin\phi}{y} - \frac{\omega_n}{K} - \left(2\zeta - \frac{\omega_n}{K}\right)\cos\phi. \qquad (44)$$

There is presently no analytical method available to solve this equation. However, the equation completely defines the slope of the solution curve $y(\phi)$ at all points of a $\phi-y$ plane, except for the points of stable and unstable equilibrium, $y=0$; $\Delta\omega/K=\sin\phi$. The limit of synchronization can thus be found graphically by starting the system with an infinitesimal velocity Δy at a point of unstable equilibrium, $y=0$; $\phi=\pi-\sin^{-1}\Delta\omega/K$, and finding the value of $\Delta\omega/K$ for which the solution curve just reaches the next point of unstable equilibrium located at $y=0$; $\phi=3\pi-\sin^{-1}\Delta\omega/K$. The method is discussed by Stoker[4] and has been used by Tellier and Preston[5] to find the pull-in range for a single time constant AFC system.

To establish the limit curve of synchronization for given values of ζ and ω_n/K, a number of solution curves have to be plotted with $\Delta\omega/K$ as parameter. The limit of pull-in range in terms of $\Delta\omega/K$ then can be interpolated to any desired degree of accuracy. The result, obtained in this manner, is shown in the dimensionless graph of Fig. 7, where $\Delta\omega/K$ is plotted as a function of ω_n/K for a damping ratio $\zeta=0.5$. Since this curve represents the stability limit of synchronization for the system, the time required to reach synchronism is infinite when starting from any point on the limit curve. The same applies to any point on the $\Delta\omega/K$-axis, with exception of the point $\Delta\omega/K=0$, since this axis describes a system having either infinite gain or zero bandwidth, and neither case has any real practical significance. The practical pull-in range of synchronization, therefore, lies inside the solid boundary. The individual points entered in Fig. 7 represent the measured pull-in curve of a particular system for which the damping ratio was maintained at $\zeta = 0.5$. For small values of ω_n/K this pull-in curve can be approximated by its circle of curvature which, as indicated by the dotted line, is tangent to the $\Delta\omega/K$-axis and whose center lies on the ω_n/K-axis. The pull-in range thus can be expressed analytically by the equation of the circle of curvature. If its radius is denoted by ζ, the circle is given by

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[3] Edouard Labin, "Theorie de la synchronization par controle de phase," *Philips Res. Rep.*, (in French); August, 1941.

[4] J. J. Stoker, "Non-linear vibrations," *Interscience;* New York, 1950.

[5] G. W. Preston and J. C. Tellier, "The Lock-in Performance of an A.F.C. Circuit," Proc. I.R.E.; February, 1953.

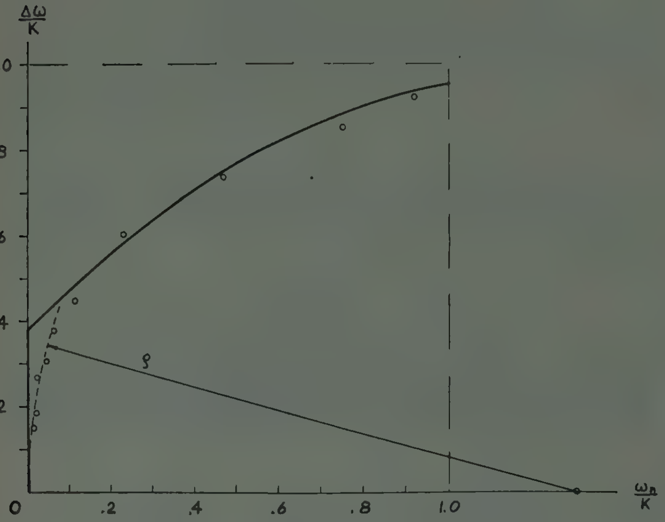

Fig. 7—Pull-in range of synchronization for $\zeta = 0.5$.

$$\left(\frac{\omega_n}{K} - \zeta\right)^2 + \left(\frac{\Delta\omega}{K}\right)^2 = \zeta^2. \qquad (45)$$

Hence, for $(\omega_n/K) \to 0$, the pull-in range of synchronization is approximately

$$\left| \Delta\omega_{\text{Pull-in}} \right|_{(\omega_n/K)\to 0} < \sqrt{2\zeta\omega_n K - \omega_n^2} \cong \sqrt{2\zeta\omega_n K}. \qquad (46)$$

ζ can be interpreted as a constant of proportionality which depends on the particular design of the system, and which increases as the system gets closer to the theoretical limit of synchronization.

Equation (46) shows that the pull-in range for small values of ω_n/K is proportional to the square root of the product of the cut-off frequency ω_n and the gain constant K. Since the bandwidth of a double time constant AFC system can be adjusted independently of the gain constant, the pull-in range of such a system can exceed the noise bandwidth by any desired amount.

V. CONCLUSIONS

The performance of an AFC system can be described by three parameters. These are the gain constant K, the damping ratio ζ and the resonance or cut-off frequency ω_n. These parameters are specified by the requirements of a particular application and define the over-all design of the system. It has been shown that among the systems with zero, single and double time constant control networks, only the latter fulfills the requirement for achieving good noise immunity, small steady-state phase error and large pull-in range.

A New Solution for the Current and Voltage Distributions on Cylindrical, Elliptical, Conical or Other Axisymmetrical Antennas*

O. ZINKE†

ABSTRACT[1]

TO FIND THE CURRENT DISTRIBUTION we only need four results derived from a treatment of the problem by means of Maxwell's equations:

1. The voltage (the scalar potential) varies nearly sinusoidally along a cylindrical antenna

$$V = V_{\max}\left[\cos 2\pi \frac{L + l_v - z}{\lambda} + j\frac{R_{\text{rad}}}{Z} \sin 2\pi \frac{L + l_v - z}{\lambda}\right]. \qquad (1)$$

L = length of the half of the antenna.
l_v = equivalent length of the end plate $\approx r_0\sqrt{2}$.
z = variable, reckoned from the plan of symmetry along the surface of the antenna
λ = wavelength.
R_{rad} = radiation resistance $= f(L/\lambda)$.
Z = characteristic impedance $= 120$ ($\log_e 2L/D$ − const.) ohms.

2. The field strength E_n normal to the antenna surface is derived from $E_n = -\partial\phi/\partial n$ as in the electrostatic case

$$E_n = -\frac{d\phi_{\text{stat}}}{dn}\left[\cos 2\pi \frac{L + l_v - z}{\lambda} + j\frac{R_{\text{rad}}}{Z} \sin 2\pi \frac{L + l_v - z}{\lambda}\right]. \qquad (2)$$

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† Zentrallaboratorium, Siemens and Halske Aktiengesellschaft, Munich, Germany.

[1] For complete paper order Document 3923 from American Documentation Institute Auxiliary Publications Project, Photo-duplication Service, Library of Congress, Washington, D. C. $2.00 for microfilm, or $3.75 for photoprints readable without optical aid.

Herein $d\phi_{stat}/dn$ is found either by field plotting or experimentally in the electrolytic trough or by calculation (method of Southwell).

3. The charge q per unit length of the antenna surface is obtained as product of the known field strength E_n and the antenna radius r_0

$$q = E_n \epsilon_0 \epsilon_{rel} 2\pi r_0. \quad (3)$$

4. By the continuity law of the current $dI_z/dz = j\omega q = j\omega E_n \epsilon_0 \epsilon_{rel} 2\pi r_0$ we thus know the slope of the current at any point z. The current I_z itself is now after integration

$$I_z = \frac{j}{60\Omega}\int_{z=L+l_v}^{z} r_0 \frac{d\phi_{stat}}{dn}\left[\cos 2\pi \frac{L+l_v-z}{\lambda} + j\frac{R_{rad}}{Z}\sin 2\pi\frac{L+l_v-z}{\lambda}\right] d\left(\frac{2\pi z}{\lambda}\right). \quad (4)$$

From (4) it is obvious that the irregularities of the function $r_0(d\phi_{stat}/dn)$ govern the distortion of the current curve. For homogeneous lines r_0 and $d\phi_{stat}/dn$ are constant; for elliptical antennas the product $r_0(d\phi_{stat}/dn)$ is constant (except the ends). In accordance with (4) the current distribution will thus remain sinusoidal.

For cylindrical antennas, however, $d\phi_{stat}/dn$ varies rapidly with z along the surface. The influence of the "distortion function" $r_0(d\phi_{stat}/dn)$ is particularly remarkable in the case of voltage resonance as seen in Fig. 1 where

(a) shows the field strength $E_{stat} = -(d\phi_{stat}/dn)$,
(b) the two voltage components V_1 and V_2 according to (1).
(c) gives the two charges q_1 and q_2 corresponding to (3).

The charge q_1 is strongly distorted near the feed point. The integration of the charge distribution curves q_1 and q_2 yields at once the currents i_1 and i_2, as shown in (d) of Fig. 1. The main current i_1 presents a maximum at the zero point of V_1 and q_1. The length from the antenna end to this maximum is nearly $\lambda/4$. It is only now that antenna shortening begins. The current i_1 becomes zero when the negative area of q_1 equals the positive area. Such is the case at $z = h'$. Now $h_m - h'$ is considerably smaller than $\lambda/4$.

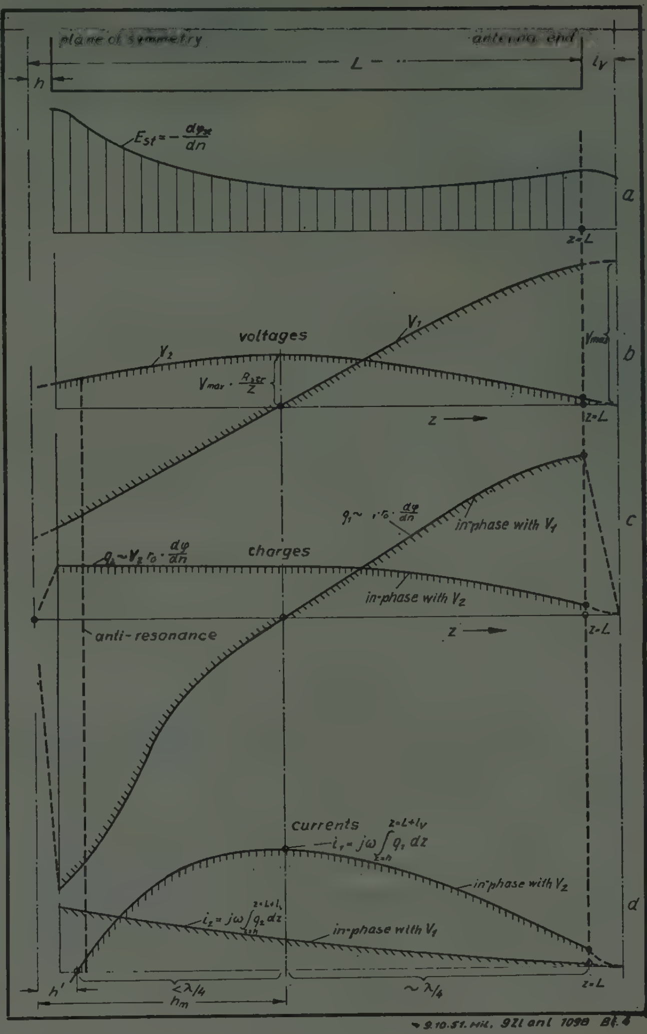
plane of symmetry
antenna end
L
h
l_v
$E_{st} = -\frac{d\varphi_{st}}{dn}$
a
z=L
voltages
V_1
V_2
$V_{max} \cdot \frac{R_{str}}{Z}$
V_{max}
b
z →
z=L
$q_1 \sim r_0 \cdot \frac{d\varphi}{dn}$
in-phase with V_1
c
$q_2 \sim V_2 r_0 \frac{d\varphi}{dn}$
charges
in-phase with V_2
anti-resonance
z →
z=L
currents
$z=L+l_v$
$-i_1 = j\omega \int_{z=h} q_1\,dz$
in-phase with V_2
d
$i_2 = j\omega \int q_2\,dz$
in-phase with V_1
h'
$<\lambda/4$
h_m
$\sim\lambda/4$
z=L
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Fig. 1—Voltage charge and current dependent on z for an antenna at voltage resonance ($L < \lambda/2$).

The shortening of the cylindrical antenna at antiresonance is thus determined by the strong-charge accumulation near the feeding gap. The reason of the nonsinusoidal current distribution is not radiation but lack of uniformity in the static surface charge. Radiation merely produces a second component in phase quadrature to the reactive component, according to (4).

* * *

Wide-Band Phase-Delay Circuit*

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Summary—**A wide-band phase-delay circuit can be constructed provided phase delay is made to depend on both differences in length and differences in cutoff frequencies of the waveguides of the circuit.**

INTRODUCTION

In UHF EQUIPMENT where signals of the same frequency are fed through two or more parallel channels it is sometimes desirable to cause a phase shift to be developed between signals in two of the channels. This might be to produce an arbitrary phase difference or to compensate for a phase shift introduced in some other part of the equipment.

Suppose that two lengths of waveguide are supplied with signals of the same frequency and phase at the input end. The relative phase at the output end will be determined by the lengths of the guides and by the respective guide wavelengths.

If the waveguides are identical except for length, a phase difference will be produced proportional to the difference in lengths. If the operating frequency is changed by accident, or by modulation, the phase shift will be changed. If, on the other hand, the waveguides are of the same length, the phase difference can be produced by making them have different guide wavelengths. In this case a change in operating frequency also will change the phase difference, but the change will now be in the opposite direction.

WIDE-BAND PHASE-DELAY CIRCUIT

It was suggested that by combining these two methods of obtaining phase delay the variation of the phase difference with frequency might be reduced, thereby producing a wide-band phase-delay circuit.

Let ϕ be the phase difference obtained in two guides of lengths L_1 and L_2.

Let f = the operating frequency.
f_1 = the cutoff frequency for guide 1.
f_2 = the cutoff frequency for guide 2.
λ, λ_1, and λ_2 the wavelengths in free space corresponding to f, f_1 and f_2.
λ_{g1} = the guide wavelength obtained in guide 1.
f_0, λ_0 = the design frequency and corresponding wavelength in free space.

Since the guide wavelength for guide 1 is

$$\lambda_{g1} = \frac{\lambda}{\sqrt{1 - \frac{\lambda^2}{\lambda_1^2}}} = \frac{c}{\sqrt{f^2 - f_1^2}}, \tag{1}$$

the total phase delay produced in guide 1 is, in radians

$$\text{phase delay} = \frac{2\pi L_1}{\lambda_{g1}} = \frac{2\pi L_1}{c}\sqrt{f^2 - f_1^2}. \tag{2}$$

A similar expression holds for guide 2; consequently the total phase difference obtained in the two guides is

$$\phi = \frac{2\pi L_2}{c}\sqrt{f^2 - f_2^2} - \frac{2\pi L_1}{c}\sqrt{f^2 - f_1^2}. \tag{3}$$

With four parameters, L_1, L_2, f_1, f_2, to be determined, one would expect to be able to impose four conditions on the design. The primary objective is to produce a requisite phase shift ϕ_0 and to maintain the phase shift as constant as possible in spite of frequency variations. The objective can be better understood by referring to Taylor's series for the phase shift

$$\phi = \phi_0 + (f - f_0)\left(\frac{d\phi}{df}\right)_0 + \frac{(f - f_0)^2}{2!}\left(\frac{d^2\phi}{df^2}\right)_0 + \cdots \tag{4}$$

To produce a very wide-band device it would be well to have as many derivatives equal zero as possible. Therefore one might try to choose the four parameters such that ϕ_0 has the required value and the first three derivatives are equal to zero. However, it is pointed out below that only the first derivative can be made zero.

Since the second and third derivatives cannot be made zero, two other specifications are required to produce a unique design. A design procedure is given below which gives a coupler having a minimum length for a required range of operating frequencies. Hence the four specifications used are

1. Specific phase difference, ϕ_0
2. Wide-Band operation; first derivative equals zero.
3. Specified frequency range.
4. Minimum length of complete coupler.

Substituting $f = f_0$ in (3) the expression for ϕ_0 is obtained

$$\phi_0 = \frac{2\pi}{c}\left[L_2\sqrt{f_0^2 - f_2^2} - L_1\sqrt{f_0^2 - f_1^2}\right]. \tag{5}$$

Introducing two new symbols

$$m_1 = \sqrt{1 - \frac{f_1^2}{f_0^2}}, \qquad m_2 = \sqrt{1 - \frac{f_2^2}{f_0^2}}, \tag{6}$$

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(5) becomes

$$\phi_0 = \frac{2\pi f_0}{c}\,[L_2 m_2 - L_1 m_1]. \tag{7}$$

The derivatives are obtained by repeated differentiation of (3) with respect of f. At $f=f_0$ the first three derivatives are written in terms of m_1 and m_2.

$$\left(\frac{d\phi}{df}\right)_0 = \frac{2\pi}{c}\left[\frac{L_2}{m_2} - \frac{L_1}{m_1}\right] \tag{8}$$

$$\left(\frac{d^2\phi}{df^2}\right)_0 = \frac{2\pi}{cf_0}\left[\frac{L_1}{m_1^3} - \frac{L_2}{m_2^3}\right] \tag{9}$$

$$\left(\frac{d^3\phi}{df^3}\right)_0 = \frac{6\pi}{cf_0^2}\left[\frac{L_1}{m_1^3} - \frac{L_2}{m_2^3} + \frac{L_2}{m_2^5} - \frac{L_1}{m_1^5}\right]. \tag{10}$$

The first derivative can be made zero by making

$$\frac{L_2}{m_2} = \frac{L_1}{m_1}. \tag{11}$$

In this case

$$\left(\frac{d^2\phi}{df^2}\right)_0 = \frac{2\pi}{cf_0}\,\frac{L_2}{m_2}\left[\frac{1}{m_1^2} - \frac{1}{m_2^2}\right] \tag{12}$$

$$\left(\frac{d^3\phi}{df^3}\right)_0 = \frac{6\pi L_2}{cf_0 m_2}\left[\frac{1}{m_1^2} - \frac{1}{m_2^2}\right]\left[1 - \frac{1}{m_1^2} - \frac{1}{m_2^2}\right]. \tag{13}$$

Clearly these cannot be made zero unless m_1 equals m_2. If use is made of (7) and (11) these derivatives can be written in terms of the phase difference desired

$$\left(\frac{d^2\phi}{df^2}\right)_0 = \frac{\phi_0}{m_1^2 m_2^2 f_0^2} \tag{14}$$

$$\left(\frac{d^3\phi}{df^3}\right)_0 = \frac{3\phi_0}{m_1^2 m_2^2 f_0^3}\left(1 - \frac{1}{m_1^2} - \frac{1}{m_2^2}\right). \tag{15}$$

By substituting (14) and (15) into (4) and rearranging slightly, the result obtained is two terms in Taylor's series for the fractional variation of phase difference with frequency

$$\frac{\phi - \phi_0}{\phi_0} = \left(\frac{f - f_0}{f_0}\right)^2 \frac{1}{2m_1^2 m_2^2} + \left(\frac{f - f_0}{f_0}\right)^3 \frac{1 - \dfrac{1}{m_1^2} - \dfrac{1}{m_2^2}}{2m_1^2 m_2^2}. \tag{16}$$

It is convenient to rewrite (7)

$$\phi_0 = \frac{2\pi}{\lambda_0}\,\frac{L_2}{m_2}\,(m_2^2 - m_1^2). \tag{17}$$

Taking $L_2/m_2 = L_1/m_1$ as constant, the maximum effect is obtained by making m_2 as large and m_1 as small as possible. Hence for a given ϕ_0 this same choice will make the length of the coupler a minimum.

Now if m_1 is small and m_2 is large, according to (6) f_1 is large and f_2 is small. In order to estimate the possible range of values of f_1 and f_2 consider a standard waveguide 0.75×0.375 inches. This has a cutoff frequency 7.87 kmc and the recommended range of operation is 10 to 15 kmc. Since

$$10 \leqq f \leqq 15$$

$$1.27 = \frac{10}{7.87} \leqq \frac{f}{f_1} \leqq \frac{15}{7.87} = 1.91, \tag{18}$$

where f_1 is the cutoff frequency of the guide. Similarly

$$1.27 \leqq \frac{f}{f_2} \leqq 1.91, \tag{19}$$

where f_2 is the cutoff frequency of guide 2.

Assume for definiteness $L_2 > L_1$ then $m_2 > m_1$ and

$$\frac{f}{f_2} > \frac{f}{f_1};$$

therefore

$$1.27 \leqq \frac{f}{f_1} < \frac{f}{f_2} \leqq 1.91. \tag{20}$$

If the range of operation is given by $f_0 \pm F$, then

$$1.27 \leqq \frac{f_0 - F}{f_1} < \frac{f_0 + F}{f_2} \leqq 1.91, \tag{21}$$

and to obtain maximum separation between f_1 and f_2 and hence between m_1 and m_2 use the equalities in (21).

$$f_1 = \frac{f_0 - F}{1.27} \tag{22}$$

$$f_2 = \frac{f_0 + F}{1.91}. \tag{23}$$

Equations (22) and (23) along with (5) and (11) give four relations to determine uniquely the four parameters L_1, L_2, f_1, and f_2.

NUMERICAL EXAMPLE

Suppose that a coupler is desired that produces a 90° phase shift, to operate between 4 and 5 kmc. Then $f_0 = 4.5$ kmc and $F = 0.5$ kmc, and (22) and (23) give

$$f_1 = \frac{4}{1.27} = 3.15 \text{ kmc} \tag{24}$$

$$f_2 = \frac{5}{1.91} = 2.62 \text{ kmc}. \tag{25}$$

The values given in (24) and (25) are extreme values. For example f_1 can be made less than 3.15 and/or f_2 can be made greater than 2.62 if desired. If there is a standard waveguide available having a cutoff frequency between 2.62 and 3.15 it can be used in place of one of the waveguides called for. However, if these extreme values are used, the over-all length of the coupler will be a minimum. If it is decided to use these extreme values then waveguides having cutoff frequencies 2.62

and 3.15 kmc should be constructed. Their cross sections will be 2.25×1.125 and 1.875×0.937 inches inside.

To determine the deviation expected compute m_1 and m_2 and substitute into (16).

$$m_1^2 = 1 - \frac{f_1^2}{f_0^2} = 0.509. \tag{26}$$

$$m_2^2 = 1 - \frac{f_2^2}{f_0^2} = 0.660. \tag{27}$$

Use

$$\frac{f - f_0}{f_0} = \pm 0.111. \tag{28}$$

Hence

$$\frac{\phi - \phi_0}{\phi_0} = \frac{0.111^2}{2(0.509)(0.660)} \pm \frac{0.111^3\left(1 - \frac{1}{0.509} - \frac{1}{0.660}\right)}{2(0.509)(0.660)} = 0.0184 \mp 0.00507. \tag{29}$$

This variation is about 2 per cent compared to a variation of 20 per cent as would be experienced if the phase difference were obtained by difference in guide length alone. In the expression above one sign corresponds to one side of the frequency band, the other sign to the opposite side of the band. Hence the design can be improved by choosing the design frequency, f_0, to one side of the central frequency. In this way the variation in phase can be made the same at both sides of the operating range. This requirement will not be investigated here.

Solving (16) for L_2,

$$L_2 = \frac{\phi_0 \lambda_0}{2\pi} \frac{m_2}{m_2^2 - m_1^2} = 20.1 \frac{\phi_0}{2\pi} \cdot \tag{30}$$

For a phase difference of one-quarter cycle

$$L_2 = \frac{20.1}{4} = 5.02 \text{ in.} \tag{31}$$

and

$$L_1 = L_2 \frac{m_1}{m_2} = 4.39 \text{ in.}$$

If it would be more convenient to have greater lengths the design can be changed to bring f_1 and f_2 nearer equality. In this illustration where the working range is 4 to 5 kmc the guides adopted have normal working ranges of 3.33 to 5 and 4 to 6 kmc. One of these could be replaced by a guide having a normal working range of, say, 3.67 to 5.5 kmc. Not only will this change increase the over-all length but it will also reduce the deviation computed in (29).

CONCLUSION

If the required range of operation is less than two-to-three this type coupler can be used. Determine the cutoff frequencies of the guides to satisfy (22) and (23). If a standard waveguide has a cutoff frequency between the cutoff frequencies computed, then the standard guide can be used and its cutoff frequency should replace one of those determined in (22) and (23) in the computations that follow. If no standard waveguide can be used, it will be necessary to have two custom-made guides instead of one. After it has been determined whether a standard guide can be used the length of the coupler is computed as in (17) or (30). If the computed length is unsatisfactory it can be increased by choosing the cutoff frequencies closer together. Generally the length computed, as in the example above, is a minimum. However, if the arbitrary selection of a standard waveguide has brought the two cutoff frequencies closer together than the values obtained by (22) and (23), the length will be greater than the minimum. In such a case it will be possible to reduce the length by discarding the standard waveguide and using two custom-made guides with a greater difference between their cutoff frequencies.

In constructing these couplers, the difference in length has been obtained by bending the guides to arcs of circles. The same theory and method of construction has also been used to obtain a "zero phase shift" angle by designing for ϕ_0 equal to 2π radians.

Coding by Feedback Methods*

B. D. SMITH†, ASSOCIATE, IRE

Summary—**This paper describes the basic feedback coder; that is, a device which converts an analog quantity such as a voltage into a digital quantity such as a binary number by the feedback method. The basic feedback coder consists of a decoding device or network together with an error amplifier and control circuits. The decoding network and its operation as a feedback element to produce a coding system are described in terms of a number system having any base, *b*. The particular case of a binary coder (base 2) is considered in detail. In addition, feedback coding systems employing a binary coded decimal number system are discussed. The advantages and disadvantages of the feedback method as compared to other coding methods are discussed. A method of nonlinear coding (or signal compression) directly in the coding process is presented.**

INTRODUCTION

THERE are many fields in which the transmission and processing of information can best be done by digital methods rather than analog methods. Pulse code modulation systems, digital computers, and digital data transmitters are examples of systems which process information in numerical form. There are many applications in which analog-to-digital conversion devices or coders are required so that the digital equipment can accept input information in analog form, such as a voltage or current. The purpose of this paper is to describe the basic feedback method of converting an electrical signal, in the form of a voltage or current, into digital form, such as a binary number.

A number of coding methods have been described in the literature. The counting method[1,2] and the coding tube method[3] have been employed in the past. The feedback method[4,5] has not received as much attention, although it offers some advantages over the counting and coding tube methods.

DECODING NETWORKS

The term "feedback" as applied to coding methods means that some form of a decoder (a device to convert from digital to analog form) is used as an integral part of the coder together with an error amplifier and control circuits. An accurate coder can be made by employing an accurate decoder as a feedback element.

The principle of operation of a decoder is based directly upon the definition of a number. Any number is represented by a series of digits, a_1, a_2, a_3, and so on, each of which can have one of a number of integral values beginning with zero. The number of digit values is the base, b, of the number system; for example, in the binary system (base 2) the digits can be either 0 or 1, and in the decimal system (base 10) they can be any integer from 0 to 9. A number is usually written with the least significant digit, a_1, last, and has a numerical value, n, defined by (1).

$$n = \sum_{k=1}^{K} a_k b^{k-1}. \tag{1}$$

For example,

$$10101 \text{ (base 2)} = 1 \times 2^4 + 0 \times 2^3 + 1 \times 2^2 + 0 \times 2^1 + 1 \times 2^0 = 21 \text{ (base 10)}.$$

If the digits of a number are represented by voltage or current sources, then it is possible to construct networks such as are shown in Figs. 1 and 2 in which the output voltage or current is related to the voltage sources in the same manner as (1). These networks act as decoders since they convert the digital information determined by the states of the sources, e_1 to e_K, into an analog quantity which is the output voltage or current. Fig. 1 shows a shunt resistor network and its equivalent circuit. The equivalent circuit is obtained by applying the superposition theorem. The open-circuit voltage, the short-circuit current, and the voltage or current into any constant load resistance are all proportional to the number

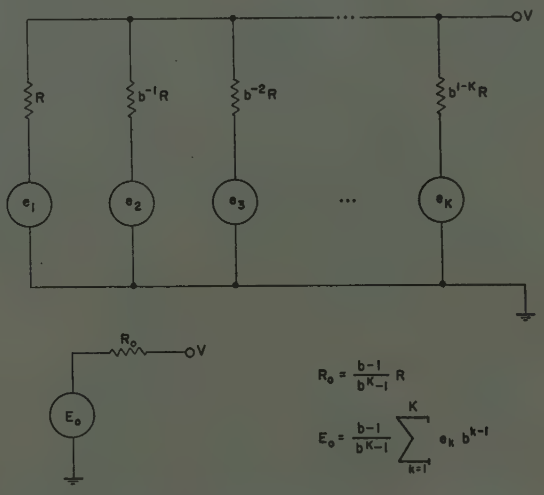
V
R
$b^{-1}R$
$b^{-2}R$
$b^{1-K}R$
e_1
e_2
e_3
e_K
R_0
V
E_0
$R_0 = \frac{b-1}{b^K-1} R$
$E_0 = \frac{b-1}{b^K-1} \sum_{k=1}^{K} e_k b^{k-1}$

Fig. 1—Shunt resistor decoding network and equivalent circuit.

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* Decimal classification: 621.375.2. Original manuscript received by the Institute, August 29, 1952; revised manuscript received February 3, 1953.

† Melpar, Inc., 452 Swann Ave., Alexandria, Va.

[1] A. H. Reeves, United States Patent Number 2,272,070 (International Standard Electric Corp.); February 3, 1942.

[2] H. S. Black, and J. O. Edson, "PCM equipment," *Elec. Eng.*, vol. 66, p. 1122; November, 1947.

[3] L. A. Meacham, and E. Peterson, "An experimental multichannel pulse code modulation system of toll quality," *Bell Sys. Tech. Jour.*, vol. 27, p. 1; January, 1948.

[4] W. M. Goodall, "Telephony by pulse code modulation," *Bell Sys. Tech. Jour.*, vol. 26, p. 395; July, 1947.

[5] B. D. Smith, "Pulse-Code Modulation Method," M.S. Thesis, M.I.T., Cambridge, Mass.; June, 1948.

represented by the voltage sources. Fig. 2 shows a ladder network having the same properties. The advantage of this network is that the impedances seen by the sources are more equal. The sources, if they are to represent the digits of a number, will have a finite number of stable states corresponding to the number base. The base 2 requires the simplest type of source since only two stable states are required. The sources in this case may be replaced by bistable devices such as flip-flops, switches, relays, and so forth. The use of a higher number base requires multistable devices such as stepping relays or combinations of bistable devices. Fig. 3 shows a binary decoding network in which the sources are replaced by relay or switch contacts and a single common "battery" E. The equivalent circuit is also shown in Fig. 3. The output voltage is proportional to the binary number represented by the contacts a_1 to a_K. If the bistable devices have internal impedance, it should be taken into account by reducing network resistors appropriately.

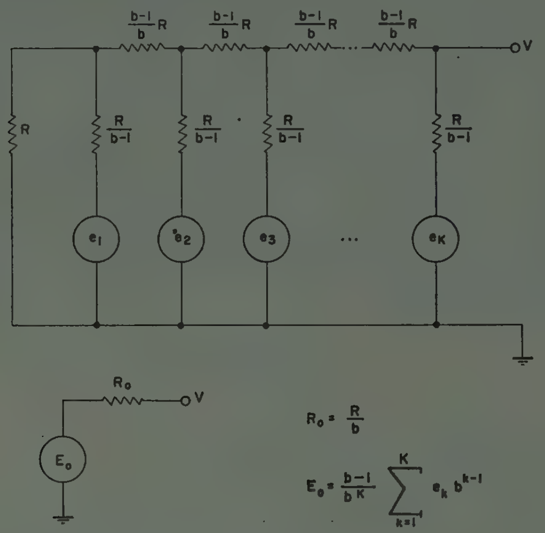

Fig. 2—Ladder resistor decoding network and equivalent circuit.

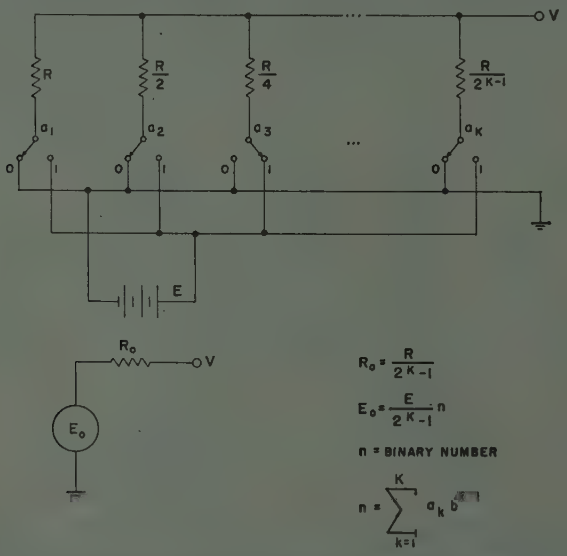

Fig. 3—Binary decoding network and equivalent circuit.

FEEDBACK CODER OPERATION

The basic feedback coder is shown in Fig. 4. The equivalent circuit of the decoding network, consisting of a voltage source proportional to the number, n, and an equivalent series resistance, is shown instead of the complete decoding network. The digits of the decoding network are controlled by the control circuits in response to the output of an error amplifier. The voltage to be coded, V_i, is connected through an input resistor, R_i, to the input of the error amplifier. The output of the decoding network is also fed to the input of the error amplifier as in a conventional feedback system. The basic coding process is to change the digits of the decoding network so as to minimize the error signal at the input to the error amplifier. The purpose of the amplifier is to indicate whether the error signal is above or below a certain potential, which for convenience will be considered zero or ground potential so that the output of the amplifier indicates the polarity of the error signal. If the error signal is above ground by any amount, the output of the amplifier will be a positive voltage; if the error signal is below ground by any amount, the output of the amplifier will be a negative voltage. When the error signal is minimized, the number represented by the digits a_1 to a_K will be the correct representation of V_i with the appropriate sign, scale factor, and zero intercept determined by the polarity, resistor values, and dc level of the decoding network.

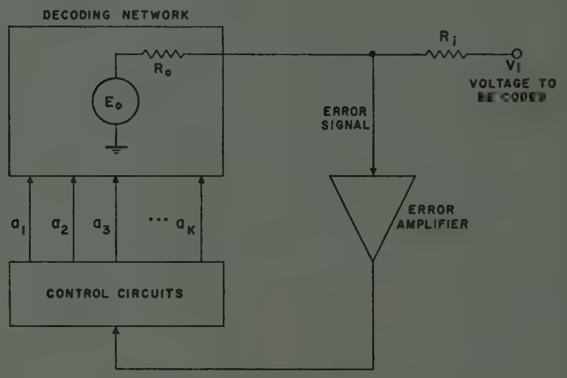

Fig. 4—Basic feedback coder.

There are two basic methods of controlling the digits in response to the error signal. One method consists of counting in unit steps only. For example, the number, n, represented by the digits can be made, 0, 1, 2, 3, and so on, until the error signal changes sign, at which time counting is stopped and the value of each digit read out.

The other method, which requires fewer operations, is to determine each digit in sequence, beginning with the most significant digit. For example, all digits may be initially reset to zero except the first (most significant) digit which is set to $b-1$, where b is the number base. If the error signal is negative, the first digit is not changed. If the error signal is positive, the first digit is changed to $b-2$, $b-3$, and so forth, until the error signal goes negative, at which point the first digit is not changed further. The second digit is then changed to $b-1$. If the error signal stays negative, the second digit is not changed. If the error signal goes positive, the second digit is changed to $b-2$, $b-3$, and so on, until the error signal goes negative, at which point the second digit is not changed further. This same process is continued until the least significant digit is measured. Another variation of this method is its complement, that is, the digits are initially reset to $b-1$ except the first which is set to 0. The first digit is changed to 1, 2, 3, and so on, until the error signal goes positive. This process is continued in the same manner to measure all of the digits. A third variation is to initially reset all digits to zero (including the first), and change the first digit to 1, 2, 3, and so forth, until the error signal goes positive. The first digit is then set back one unit, and the second digit is measured in the same manner. In any case a maximum of $K(b-1)$ operations are required as compared to b^K operations with the "counting" method.

As mentioned previously, the digits of a number having a base greater than 2 can be represented by combinations of bistable devices. An example is the binary-coded decimal system in which a decimal digit is represented by four binary digits. In this case a single decimal digit in the decoding network would be replaced by four bistable sources and four weighting resistors. The coding process may be conducted as given above, or it may be conducted as if the coder were actually a binary coder, that is, the value of each binary digit representing a decimal digit may be determined in sequence. Any general coded b-base system may be employed in this manner provided each element of the particular code has a fixed weighting factor. For example, a coded decimal system employing a four-element code to represent a decimal number in which the weights of the elements are 5, 4, 2, 1, respectively, can be used. In this case the coding process would begin with all digits reset to 0 except the first which is set to 5. If the error signal is negative, the 5 is retained. If the error signal is positive, the element representing 5 is reset to zero. In the next step a 4 is added, and the error signal indicates whether or not the element representing 4 should be retained or reset to zero. This process is continued for the remaining elements of the first decimal digit so that one decimal digit is measured in four steps. The same process is continued for the remaining decimal digits. It should be noted that a four-element code for a decimal digit results in six unused combinations since 16 combinations of four elements are possible. The coder will not generate any of these unused combinations provided the input voltage is within the range for which the coder was designed.

BINARY FEEDBACK CODER

The description of the coding process above is for the general case of any number base, b. For a binary coder the coding system is the simplest. Fig. 5 shows one type of binary feedback coder in block diagram form. In this figure the digits are represented by flip-flops, and the control circuits consist of one gate circuit for each digit except the last. For convenience, the order of numbering the flip-flops and the like is reversed, so that subscript 1 corresponds to the *most* significant digit since this digit is determined first. A source of timing pulses, X_1 through X_K, is required. These pulses are phased in time so that X_1 occurs first, X_2 second, X_3 third, and so on, as shown in Fig. 6. Initially, all flip-flops are set to zero by pulse X_1 except $FF1$ which is set to 1. If the error signal is negative, then $FF1$ is not changed further. If the error signal is positive, then $FF1$ is reset to 0 by means of $GT1$ when pulse X_2 occurs. $FF2$ is set to 1 by the same pulse X_2, and if the error signal is then nega-

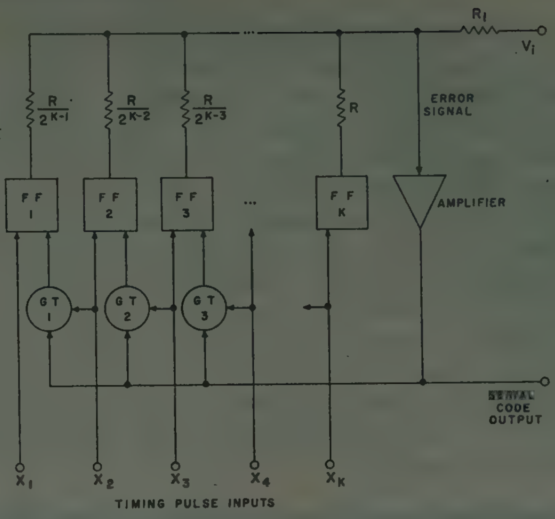

Fig. 5—Binary feedback coder.

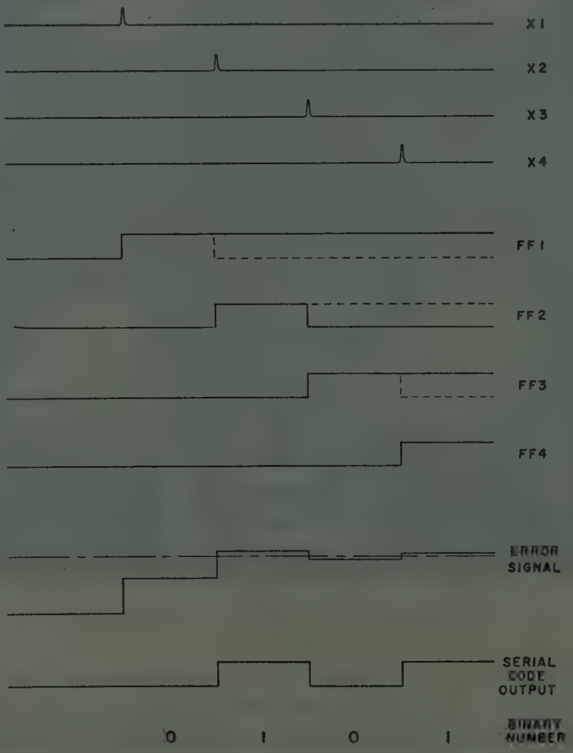

Fig. 6—Waveforms of binary coder shown in Fig. 5.

tive, *FF*2 is not changed further. If the error signal is positive, then *FF*2 is reset to 0 by means of *GT*2 when pulse X_3 occurs. *FF*3 is set to 1 by pulse X_3, and so on. The same process is repeated through all of the digits. Fig. 6 shows typical waveforms of the timing pulses, the flip-flop outputs, the error signal, and the serial code output (error amplifier output) for a particular value of the input voltage, V_i. The dotted lines show other possible waveforms produced by the flip-flops for other values of V_i. The error signal waveform shows that the error signal changes at each step in the coding process so as to approach zero. There need be no lost time between individual coding processes, that is, the first digit of one code can follow immediately after the last digit of the preceding code.

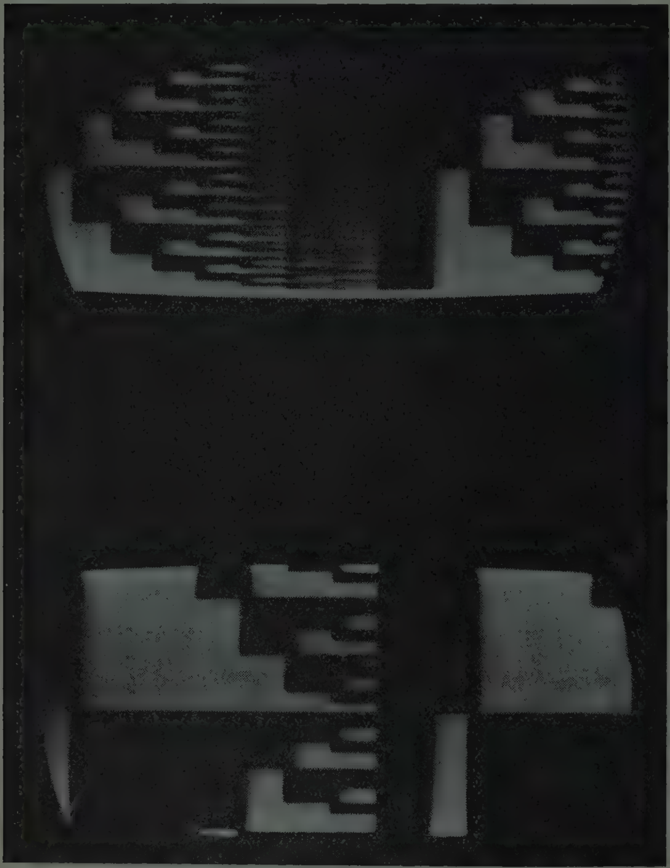

Fig. 7—Modulation pattern of an eight-digit coder.

Fig. 7 shows two views of the modulation pattern of an 8-digit binary feedback coder. This pattern is obtained by applying a slowly varying voltage simultaneously to the coder and to the *Y*-axis of an oscilloscope. The oscilloscope beam is swept horizontally in synchronization with the coder repetition rate so that a raster similar to a television raster is produced. When the *Z*-axis (intensity) is modulated by the output pulse code, the characteristic patterns shown in Fig. 7 are obtained. The upper view shows the complete pattern obtained with the slowly varying signal covering the full range of the coder (256 levels). The lower view shows the same pattern with the signal fed to the coder attenuated so as to cover only 20 levels. This results in an "expanded" view showing the details of the last digit which are unresolved in the upper view. Both patterns cover (horizontally) one complete code group and five digits of the following code group. The time scale is 4.1 μsec per digit. There is an unused time interval following each code to provide sufficient sampling time since this particular coder is used in a multiplex system. The "expanded" view of the modulation pattern shows that the eighth digit is reasonably uniform and free of noise. This type of modulation pattern is useful in aligning the coder and detecting errors in the operation of the coder.

SERIAL CODE OUTPUT

It should be noted that the digital information obtained from the coding process is obtained from the output of the error amplifier. In a binary coder, for example, the error amplifier output is a serial binary pulse code, as shown in Fig. 6, which is suitable for transmission to a remote point or for use in a digital computer. The serial code output is a binary number proportional to the input voltage, V_i. The state of the flip-flops at the end of the coding period is actually the complement of this number for the particular circuit and polarities assumed in Figs. 5 and 6. There is no need to read out or otherwise sense the condition of the flip-flops unless all digits must be transmitted in parallel or in some unusual manner. The most significant digit always occurs first in time in the serial output of a feedback coder.

In the general *b*-base coder the error amplifier output is also a serial pulse code of a particular type depending upon how the coding is conducted. If the coding process is conducted by counting in unit steps, then the serial output is a pulse whose duration is variable in unit steps; that is, a 0 is represented by a pulse of zero duration, a 1 by a pulse of 1 unit duration, a 2 by a pulse of 2 units duration, and so on. It is of passing interest to note that if a binary coder is considered a special case ($b=2$) of the general *b*-base coder, then the binary code output might be considered as a series of pulses of 0 or 1 unit *duration* and not a series of pulses of 0 or 1 unit *amplitude*. If a coded *b*-base number system is employed and the coding process is conducted as if the coder were straight binary, then the error amplifier output is a serial code in which each *b*-base digit is represented by a serial pulse code. For example, in a two-digit decimal coder using a code in which each decimal digit is composed of four elements, the serial output would be an eight pulse code. The first four pulses would be the coded decimal representation of the first decimal digit, and the last four elements the representation of the last or least significant decimal digit. A feedback coder operates so that the digits and also the elements of each digit are produced in the order of their significance, that is, the most significant element of the most significant digit first, and so forth. The so-called "excess−3" decimal code (that is, 0=0011, 1=0100, 2 = 0101, and so on) can be generated directly provided each decimal register in the coder is reset to 0011 instead of 0000 and set to 1000 when the first element of that particular decimal digit is to be found. Otherwise the control circuits are the same as a binary coder.

COMPARISON WITH OTHER CODING METHODS

The relative merits of the feedback coding method in comparison with the counting and coding tube methods depend, of course, upon the particular application; but certain general advantages and disadvantages can be given. With respect to the counting method of binary coding, these points of comparison are important:

1. With the counting method it is necessary to operate two coders alternately or provide a temporary storage register if no lost time can be tolerated in a serial output coder. This is due to the fact that a single counter cannot be read out at the same time it is counting. With a feedback coder the first digit of one code group can follow immediately after the last digit of the preceding code group with no lost time.
2. Serial output is obtained very simply in a feedback coder. With respect to parallel output, the two methods are approximately equal in complexity.
3. With respect to speed, the feedback coder can be made considerably faster because of the fewer operations required.
4. With respect to accuracy and reliability, the two methods are approximately equal. Since the bistable elements in a feedback coder are either set or reset instead of being triggered as in a flip-flop counter chain, better reliability can be attained. However, there are sources of errors in a feedback coder (such as inaccuracies in the decoding network resistors) which do not occur with the counting method.

In a comparison of the feedback method with the coding tube method, the following points are important:

1. The coding tube method requires separate circuits for quantization since large coding errors can occur if the coding tube input changes even a small amount during the coding process. With the feedback coder no such circuits are required; if the signal changes during the coding process, the generated code always corresponds to some value of the input signal in between the values at the beginning and end of the coding period.
2. With the feedback method it is not necessary to use special nonbinary codes, such as the Grey or reflected binary code, to eliminate errors such as are produced when the input signal is on the dividing lines between levels in a coding tube. This source of error is eliminated by the internal positive feedback or "flip-flop" action designed into the error amplifier of a feedback coder.
3. With respect to speed, the coding tube method is definitely superior. A coding tube system has been developed which is suitable for coding wide-band signals such as television, indicating a coding speed which is approximately 50 times that of the feedback coder whose modulation pattern is shown in Fig. 7.
4. In accuracy, the two methods are comparable.
5. The coding tube method makes it possible to transmit a binary code with the least significant digit first, which permits the use of simpler decoding devices, such as the Shannon-Rack decoder.

The choice of one or the other of these coding methods depends upon all of the factors involved, such as speed, accuracy, size, weight, ruggedness, and the requirements of the over-all system in which it is to be employed.

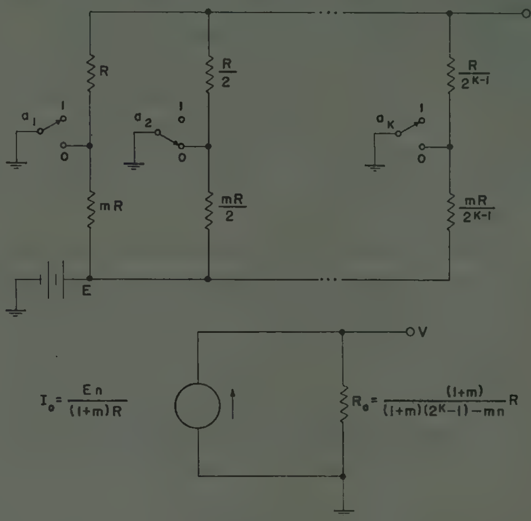

Fig. 8—Nonlinear decoding network and equivalent circuit.

NONLINEAR CODING METHODS

The feedback coding method offers the possibility of nonlinear coding or signal compression in the coding process. Fig. 8 shows a binary decoding network which is nonlinear in that the impedance seen at point V is a function of the settings of the switches. The equivalent circuit in terms of a current source and equivalent resistance is also shown in Fig. 8. The short-circuit current is linear with n, but the open-circuit voltage is proportional to n divided by a constant minus n. That is, as the switches are changed from one number to the next, the open-circuit voltage will lie on one of the curves

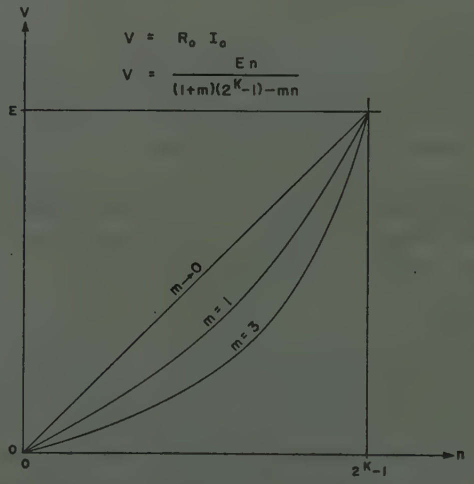

Fig. 9—Characteristic curves of the decoding network shown in Fig. 8.

(section of hyperbolas) shown in Fig. 9, depending upon the value of the constant m. A decoder of this type produces a voltage which is an expanded (positive curvature) function of the binary number, n. By using this type decoder in the feedback loop of a feedback coder, n will be a compressed (negative curvature) function of the input voltage to the coder. A cathode follower can be used to preserve the open-circuit condition for the network. The over-all transfer function from the voltage input to a coder, to the output voltage of a decoder in a complete system will be linear. The effect of the compression-expansion process is that signals, such as speech, which require a wide dynamic range can be transmitted with less average quantization distortion. Symmetrical compression and expansion curves can be obtained by modifying the coding and decoding processes for plus-minus operation.

In comparison with other compression systems, such as are obtained by the use of varistors or other nonlinear impedances, this method has the advantage of possibly better over-all system linearity without the need for closely matching pairs of varistors. The compression curves are mathematically defined and can be changed readily if desired. Simple networks producing smooth nonlinear curves other than sections of hyperbolas should be possible but have not been found. The nonlinear codes represent an additional step toward the ideal coding system prescribed by information theory which takes into account all *a priori* information on the inputs, outputs, and noise in a communication system.

CONCLUSIONS

The problem of analog-to-digital conversion can be solved by a number of coding systems, one of which is the feedback system described in this paper. A feedback coder can be made as accurate as coders using other methods, and in many applications offers some advantages. The feedback coder can be designed to operate with any number base and with the coded-decimal systems used in computers, but is simplest when employed as a binary coder to generate a serial pulse code. The feedback coder can be employed to generate a number which is a nonlinear function of the input signal, suitable for use where signal compression is desired.

ACKNOWLEDGMENT

The author wishes to express his appreciation for the advice and encouragement of Dr. W. G. Tuller, and for the help of B. H. Dennison and A. H. Ballard, all of the Melpar engineering staff.

Discussion on

"Synthesis of Narrow-Band Direct-Coupled Filters"*

H. J. RIBLET

J. Reed:[1] Dr. Riblet's recent paper is based on the simple but erroneous theory that one is dealing with constant shunt susceptances. His formula (7) can readily be shown to be identical with that of Mumford's.[2] Since the contribution[3] of the frequency consciousness of inductive irises to the Q of filter stages like those Dr. Riblet considers is of the order of 10 per cent, it would seem that this contribution is definitely not a second-order effect.

H. J. Riblet: I am indebted to John Reed for drawing my attention to the applicability of the results of his paper entitled, "Low Q Microwave Filters" to the design of narrow-band filters. In my examination of the literature, I was misled by the title of his paper. The numerical equivalence between my expression for the Q of a cavity, equation (7), and the formulas used by Mumford was shown in graphs submitted with the first version of the paper and then omitted finally in the interest of brevity. Mr. Reed is correct in his observation that equation (7) and Mumford's $Q \doteq \sqrt{B^4+4B^2}\tan^{-1}(2/B)/4$ are identical. Those who wonder why my exact expression (7) and the approximate expression used by Mumford are identical will find the answer in a slight difference in the definition of Q.

Whether or not the 10-per cent error, referred to by Reed, in the Q's of the end cavities of "maximally flat" filters is significant in the design of narrow-band filters is open to question in view of the approximate nature of the entire procedure and the considerable experimental success which previous workers have had using the assumption of constant susceptance. Nevertheless, the inclusion of this frequency dependence does appear to be sound; and since it involves no additional design effort, it certainly merits serious consideration.

For this purpose, the method of my paper is immediately applicable. In proceeding from equation (2) to the first-order admittance transformation of the antiresonant element analogous to equation (4), one only has to include the variation of ϕ with respect to Ω. For inductances, $B=-A\lambda g$, and it is readily found that $d\phi/d\Omega=\sin\phi_0\cos\phi_0$, where $\phi_0+(2\pi l/\lambda g_0)=\psi_0=0$ at antiresonance. For capacitances, $B=A/\lambda g$ and $(d\phi/d\Omega)=-\sin\phi_0\cos\phi_0$ at antiresonance. Then we find that

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* H. J. Riblet, "Synthesis of narrow-band direct-coupled filters," PROC. I.R.E., vol. 40, pp. 1219–1223; October, 1952.

[1] Raytheon Manufacturing Co., Newton, Mass.

[2] W. W. Mumford, "Maximally flat filters in waveguide," *Bell Sys. Tech. Jour.*, vol. XXVII, p. 699, Fig. 11; October, 1948.

[3] J. Reed, "Low Q microwave filters," PROC. I.R.E., vol. 38, p. 794, Formula 8 and Table I; July, 1950.

$$Y_i=\frac{j(1+\cos\phi_0-k_1\Omega)+k_2\Omega Y_0}{k_2\Omega+j(1-\cos\phi_0+k_1\Omega)}\tag{4'}$$

where

$$k_1=\pm\sin^2\phi_0\cdot\cos\phi_0$$
$$k_2=\phi_0\mp\sin\phi_0\cdot\cos\phi_0,$$

and the upper signs are used with inductances and the lower signs with capacitances. A pair of such antiresonant elements separated by a half-guide wavelength has the first-order admittance transformation,

$$Y_i=\frac{j\dfrac{(1+\cos\phi_0)(\pi(1+\cos\phi_0)-2k_2)\Omega}{(1+\cos\phi_0)(1-\cos\phi_0)+2k_1\cos\phi_0\Omega}+Y_0}{1+j\dfrac{(1-\cos\phi_0)(\pi(1-\cos\phi_0)-2k_2)\Omega}{(1+\cos\phi_0)(1-\cos\phi_0)+2k_1\cos\phi_0\Omega}Y_0}.\tag{5'}$$

For frequencies in the pass band of the filter $Y_0\approx 1$, and we may expand by the binomial theorem and obtain

$$Q=\frac{\cos\phi_0(\pi-k_2)}{\sin^2\phi_0}.\tag{7'}$$

For the inductive case this is identically equivalent to Mr. Reed's equation (8).

The extension to the design of direct-coupled filters proceeds as in my paper. Analogous to (9) we find

$$Y_i=j\frac{(1-\cos\phi_0+k_1\Omega)+(k_2-\pi)\Omega Y_0}{(k_2-\pi)\Omega+j(1+\cos\phi_0-k_1\Omega)Y_0},\tag{9'}$$

and for (10)

$$Y_i=j\frac{[(1-\cos\phi_0')(1-\cos\phi_0'')+A\Omega]+B\Omega Y_0}{C\Omega+j[(1+\cos\phi_0')(1+\cos\phi_0'')-D\Omega]Y_0},\tag{10'}$$

where

$$A=k_1''(1-\cos\phi_0')+k_1'(1-\cos\phi_0'').$$
$$B=k_2'+k_2''+(\pi/2-k_2'')\cos\phi_0'-(\pi/2-k_2')\cos\phi_0''$$
$$-\frac{3\pi}{2}-\pi/2\cos\phi_0'\cos\phi_0''.$$
$$C=k_2'+k_2''-(\pi/2-k_2'')\cos\phi_0'+(\pi/2-k_2')\cos\phi_0''$$
$$-\frac{3\pi}{2}-\frac{\pi}{2}\cos\phi_0'\cos\phi_0''.$$
$$D=k_1'(1+\cos\phi_0'')+k_1''(1+\cos\phi_0').$$

As before, equivalence at resonance is established by equation (11). Moreover, an approximate first-order equivalence can be obtained by a suitable transformation of the Ω scale, for any given numerical case.

Correspondence

A Note on Sommerfeld's 1909 Paper*

Sommerfeld presented, in his 1909 paper,[1] the development of approximate expressions for the field strength due to a vertical oscillating doublet located at the surface of a plane earth with arbitrary electrical properties. Sommerfeld's formulas, however, have been found to be in error. Though Niessen discussed the mathematical error appearing in Sommerfeld's derivation in a 1937 article[2] (in German), the contents of this paper are not very widely known. Thus, articles still appear rather frequently in the literature giving erroneous explanations as to why Sommerfeld's results are in error, two of the more notable of these being the papers by Kahan and Eckart[3] and Epstein.[4]

The purpose of this note is to point out that, as explained by Niessen, the error is due to an incorrect choice of a square root that appears in Sommerfeld's variable, α. The angle of α^2 falls in the first quadrant, and Sommerfeld made the natural but erroneous choice of placing α in the first quadrant. A closer analysis reveals that the proper angle to be associated with α is in the third quadrant.

The critical portion of Sommerfeld's derivation involves the approximating of an integral where the path of integration is along the positive real axis. This integral is converted into one whose path is around a branch cut and a pole. The contributions to the integral associated with the portion of the path around the branch cut and around the pole will be denoted by Q and P, respectively. Sommerfeld arrived at the same results by two methods, the first involving evaluating Q and P independently by integrating around the branch cut and pole separately, and the second involving determination of the value of $(Q+P)$ by employing a single path inclosing both the branch cut and the pole. Sommerfeld's formulas have been found to be in error by an amount equal to P.

Some of the shortcomings of the explanations given in references 3 and 4 as to why Sommerfeld's results are incorrect are given briefly below. Epstein notes that P satisfies the wave equation and, indeed, all the conditions of a solution, except that it is singular at $r=0$ for all z and not just at $z=0$, as required by the problem. Thus Sommerfeld's initial integral satisfies the same conditions when the path of integration is along the positive real axis and when the path is transformed to pass around the other side of the pole, except for the type of singularity at $r=0$. Epstein interprets this as showing that Sommerfeld's uniqueness proof is not valid and that the transformed path must indeed be the correct one, since this changes Sommerfeld's results by P, the amount they are in error. Actually Sommerfeld embodies the type of singularity at $r=0$ into his uniqueness proof which is perfectly valid. A quick and easy check of Sommerfeld's integral with the path along the positive real axis reveals that it is singular at $r=0$ for $z=0$ only and thus is the correct path to be taken in this problem rather than the path suggested by Epstein.

Kahan and Eckart point out that a saddle point is located between the pole and the branch cut. If the value of the residue of the pole is thought of as being calculated by the saddle point method, the major contribution comes from that portion of the path in the neighborhood of the saddle point. They reason, on the basis of the path of integration encircling the branch cut and pole separately, that there will be a contribution to Q when the integration path passes the saddle point (but in the opposite direction) that will just cancel P and that Sommerfeld must have overlooked the existence of the saddle point and so does not properly take it into account. However, when considering this path of integration Sommerfeld obtained an asymptotic series and replaced it by a function having the same asymptotic expansion. He carefully pointed out that, simply due to the character of asymptotic expansions, his results could be in error by an amount of the order of magnitude of P, the consideration of the saddle point thus being of no importance. Sommerfeld actually considers the single path inclosing both the branch cut and pole when developing the convergent series expansion, which rigorously verifies his final results. Thus Kahan and Eckart's comments are not relevant.

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* Received by the Institute, March 30, 1953. Prepared at Cornell University under U. S. Air Force Contract AF 33(038)-1091. Now with Electrical Engineering Research Laboratory, The University of Texas, Austin, Texas.

[1] Sommerfeld, *Ann. Physik*, 28, p. 665; 1909.

[2] Niessen, *Ann. Physik*, 29, p. 585; 1937.

[3] Kahnn and Eckart, PROC. I.R.E., 38, p. 807; 1950.

[4] Epstein, *Proc. Nat. Acad. Sci.*, 33, p. 195; 1947.

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Collector-Base Impedance of a Junction Transistor*

Recently, J. M. Early[1] described an extension of Shockley's theory for the current voltage relationships of a junction transistor which explained certain experimental results. For example, Early's analysis included an explanation for observed values of collector resistance.

It is possible to extend Early's analysis to include the effect of the frequency of the ac variations. As a result of this further extension, an experimentally observed frequency variation of the open-circuit collector-base admittance of fused-impurity *p-n-p* junction transistors has been explained.[2] The nature of this frequency variation may be described moderately well by an (approximate) expression of the form[3]

$$1/z_{22}^{(b)} = j\omega C_c + g_c + j\omega C_d\left[\tanh (j\omega\tau_D)^{1/2}/(j\omega\tau_D)^{1/2}\right], \qquad (1)$$

where it is assumed that the internal base impedance is small in comparison with the right-hand member of (1). In this equation, $z_{22}^{(b)}$ is the open-circuit collector-base impedance; C_c is the barrier capacitance of the collector-base diode; g_c is the low-frequency collector conductance, a theoretical value for which was calculated by Early; C_d is a diffusion capacitance due to charge stored in the base-layer of the transistor by the dc emitter current; and τ_D is a constant that may be determined from the alpha-cutoff frequency[4] $\omega_c/2\pi$ of the transistor from the relation $\tau_D = 2.43/\omega_c$. The diffusion capacitance C_d may be determined from the difference between the value of the low-frequency collector-base capacitance corresponding to a given emitter current and that for zero emitter current.

Alternatively, the nature of the frequency variation of $z_{22}^{(b)}$, as well as that of the current-amplification factor α may be described *approximately* by a new equivalent circuit employing an RC transmission line, as shown in Fig. 1. In this circuit, z_e, z_b are complex emitter and base impedances, μ is a complex form of the voltage-feedback factor introduced by Early, α_0, g_c and C_c are as defined above, and r_d and c_d are the per-unit-length constants of the line of length w, i.e., $r_d = R_d/w$, $c_d = C_d/w$, with C_d as defined above, and with $R_d \equiv \tau_D/C_d$.

At low frequencies, the open-circuit collector-base admittance reduces to that of a parallel combination of the conductance g_c, the collector capacitance C_c, *and the additional diffusion capacitance* C_d. As the frequency is increased, the open-circuit collector-base conductance increases with increasing frequency, ultimately as $\omega^{1/2}$, whereas the capacitance decreases with increasing frequency as $\omega^{-1/2}$ from its low-frequency value of $(C_c + C_d)$ to an ultimate value of C_c.

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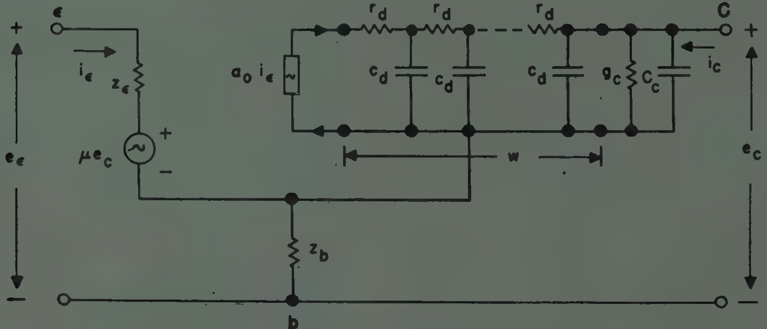

Fig. 1—Equivalent circuit including RC transmission line for junction transistor.

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* Received by the Institute, December 22, 1952.

[1] J. M. Early, "Effects of space-charge layer widening in junction transistors," Proc. I.R.E., vol. 40, pp. 1401–1406; November, 1952.

[2] Details of the extension and the experimental results will be described by the writer in a forthcoming paper.

[3] This expression follows from the theoretical result if it is assumed that the low-frequency value α_0 of the current-amplification factor is close to unity.

[4] Alpha-cutoff frequency $\omega_c/2\pi$ is defined as the frequency for which the amplitude of the current-amplification factor α is 3 db below its low-frequency value α_0. According to Shockley's theory (W. Shockley, *et al.*, "*p-n* junction transistors," *Phys. Rev.*, vol. 83, no. 1, p. 161; July 1, 1951), the frequency variation of α under certain conditions may be described by the expression $\alpha_0/\cosh(j\omega\tau_D)^{1/2}$. Hence, when $\omega\tau_D = 2.43$, $|\alpha| = 0.707\ \alpha_0$. For further details, see R. L. Pritchard, "Frequency variations of current-amplification factor for junction transistors," Proc. I.R.E., vol. 40, appendix, pp. 1480–1481; November, 1952

Remarks on the Neutrino Communication System*

There have recently appeared in these pages two communications dealing with the possible role played by the neutrino in telepathic communication. Hammond[1] first suggested that the neutrino may be the agency for such communication, and based his suggestion primarily upon the fact that the interaction of neutrinos with matter is extremely small; so small that barriers would not interfere with telepathic transmission. Following this suggestion, Tyson[2] noted that the potassium isotope of mass 40, which is radioactive, is found in human organisms and would be expected to emit neutrinos during its decay process. This writer pointed out further that the amount of K^{40} in the human brain would result in the emission of some 400 neutrinos per second.

Apart from any considerations of the evidence for or against telepathic communication, it should be evident that the neutrino can play no part in such a phenomenon. In the first place, the argument set forth by Hammond regarding the great penetrating power of the neutrino applies as well to the interaction of this particle with the brain. The detection of a neutrino signal implies the transfer of energy to the receiver, while the fact is that the neutrino has not yet been observed directly simply because its interaction with matter is so very weak. Experiments to date have succeeded only in establishing upper limits to the cross-sections, the limits being imposed by the sensitivity of the detection equipment. These measurements have shown that the scattering cross-section[3] is certainly less than 10^{-34} cm^2/electron, while the cross-section for the expected process of inverse β-decay is estimated[4] to be of the order of 10^{-44} cm^2. Unless there is some process, entirely outside our present knowledge, through which a neutrino can transfer energy to a free electron with good probability, the weak interaction argues most strongly against its participation in thought transmission.

In addition, one must consider the question of noise in the communication channel. All radioactive sources which decay via a beta process emit neutrinos with energies ranging from zero to several Mev in some cases. For example, the C^{14} present in the body emits some $2 \cdot 10^3$ neutrinos/sec, and the radium in a typical luminous dial watch produces about 10^6 neutrinos/sec. While the neutrino emission from such sources, and from the natural radioactivity in the earth, is considerably greater than that from the K^{40} in the brain, even these sources are negligible in comparison with the emission from the sun. This is so great that the neutrino intensity at the earth is estimated[5] to be 10^{11} neutrinos/sec-cm^2. Nor would this be the only major neutrino flux on the earth. In the vicinity of a high power reactor one could expect a flux of the order of 10^{12} neutrinos/sec-cm², and even at a distance of 10 miles from the explosion of a nominal atomic bomb, the flux at the end of the first minute would be some 10^9 neutrinos/sec-cm². The transmission of a neutrino signal in the presence of such a high noise level is inconceivable.

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* Received by the Institute, April 8, 1953.

[1] A. L. Hammond, "Note on Telephathic Communication," Proc. I.R.E., vol. 40, p. 605; May, 1952.

[2] G. N. Tyson, "On the possibilities of a neutrino communication system," Proc. I.R.E., vol. 41, p. 294; February, 1953.

[3] J. H. Barrett, *Phys. Rev.*, vol. 79, p. 907; 1950.

[4] H. A. Bethe, "Elementary Nuclear Theory," John Wiley and Sons, Inc., New York; 1947.

[5] E. Fermi, "Nuclear Physics," *Rev. Ed.*, U. of Chicago Press, Chicago; 1950.

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Response Characteristics*

"The Response Characteristics of Resistance-Reactance Ladder Networks," treated by R. R. Kenyon in the May issue of the Proceedings of the I.R.E., can be obtained with much greater facility by means of the methods presented in my paper on the same type of network.[1]

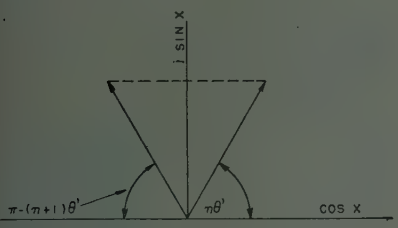

Fig. 1.—Positioning of $e^{jn\theta'}$ to satisfy $\sin(n+1)\theta' = \sin n\theta'$.

It has been shown there (Eq. j) that for the n-section network of Fig. 1:

$$\left.\frac{E_0}{E_i}\right|_n = \frac{\sinh\theta}{\sinh(n+1)\theta - \sinh n\theta} \qquad \text{(j)}$$

$$\cosh\theta = 1 + \frac{pT}{2}$$

is equivalent to Kenyon's or Tschudi's[2]

$$\left.\frac{E_0}{E_i}\right|_n = \frac{1}{\sum_{m=0}^{n} a_{m,n}(pT)^m}$$

$$a_{m,n} = \binom{n+m}{2m},$$

where the variables p and s are interchangeable.

The partial-fraction expansion of the voltage transfer ratio (j)

$$\left.\frac{E_0}{E_i}\right|_n = \frac{1}{\sum_{m=0}^{n} a_{m,n}(pT)^m} = \frac{1}{\prod_{m=1}^{n}(pT - b_{m,n})} = \sum_{m=1}^{n}\frac{B_{m,n}}{pT - b_{m,n}}$$

is required in connection with transient problems. In order to obtain it, the roots $b_{m,n}$[3] of $\sum_{m=0}^{n} a_{m,n}(pT)^m = 0$ must be found. Rather than solve this n'th order equation, consider the equivalent equation

$$\sinh(n+1)\theta = \sinh n\theta.$$

It is satisfied by purely imaginary values of $\theta(\theta = j\theta')$, such that

$$\pi - (n+1)\theta' = n\theta',$$

as is apparent from Fig. I. Consequently, the complete solution is

$$\theta' = (2m-1)\frac{\pi}{2n+1} = \alpha_m, \qquad m = 1, 2, 3, \cdots, n.$$

Since $pT = 2(\cosh\theta - 1)$, the n roots $pT = b_{m,n}$ are given by

$$b_{m,n} = 2(\cos\alpha_m - 1).$$

While Kenyon's method requires the solution of an n'th order equation, n being the number of network sections, the present method reduces the process to looking up the values in a table of cosines. The simplification is very substantial when several sections are involved.

Furthermore, even the coefficients of the partial fractions can be compared with greater economy by means of this method. Since the roots are all linear and distinct, the coefficients $B_{m,n}$ are given by

$$B_{m,n} = \left.\frac{\sinh\theta}{\frac{1}{T}\cdot\frac{d}{dp}[\sinh(n+1)\theta - \sinh n\theta]}\right|_{\theta=j\alpha m} = \frac{1-\cos 2\alpha_m}{(2n+1)\cos n\alpha_m}.$$

If the input $e_{i(t)}$ is of the unit impulse form (Dirac function), so that the Laplace transform $E_{i(p)} = 1$, then the output voltage as a function of time is given directly and without intermediate calculations by

$$e_{0(t)} = \frac{e^{-2t/T}}{(2n+1)T}\sum_{m=1}^{n}\frac{1-\cos 2\alpha_m}{\cos n\alpha_m}e^{2\cos\alpha_m\cdot t/T}$$

$$\alpha_m = (2m-1)\frac{\pi}{2n+1}, \quad m = 1, 2, 3, \cdots, n.$$

Supposing the input voltage $e_{i(t)}$ to be a unit step function (Heaviside function), the partial fraction $1/p$ is added and the coefficients of the other partial fractions become

$$T\frac{B_{m,n}}{b_{m,n}} = -\frac{1+\cos\alpha_m}{(2n+1)\cos n\alpha_m}\cdot T.$$

Consequently, the output voltage as a function of time is given explicitly by

$$e_{0(t)} = 1 - \frac{e^{-2t/T}}{2n+1}\sum_{m=1}^{n}\frac{1+\cos\alpha_m}{\cos n\alpha_m}e^{2\cos\alpha_m\cdot t/T}$$

$$\alpha_m = (2m-1)\frac{\pi}{2n+1}, \quad m = 1, 2\ 3, \cdots, n.$$

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* Received by the Institute, May 23, 1951.

[1] Leo Storch, "The Multisection RC Filter Network Problems," Proc. I.R.E., pp 1456–1458; November, 1951.

[2] R. R. Kenyon, "Response characteristics of resistance-reactance ladder networks," Proc. I.R.E., vol. 39, pp. 557–559; May, 1951.
E. W. Tschudi, "Admittance and Transfer Function for a n-Mesh RC Filter Network," Proc. I.R.E., vol. 38, pp. 309–310; March, 1950.

[3] This $b_{m,n}$ and the one used in footnote reference 1 have entirely different meanings.

A Note on the Impedance Transformation Properties of the Folded Dipole*

The impedance transformation properties of the half wave folded dipole (Fig. 1) have been investigated by R. Guertler[1] under the following restrictions: (a) The cross-section dimensions of the elements and the distance between them is small in comparison with λ/2. (b) The elements are of circular (or nearly circular) cross section. (c) $[a_2/a_1 > 1;\ s/a_2 \geqq 2{,}5$ (Fig. 1)]. It is the purpose of this note to point out a way of removing restrictions (b)(c).

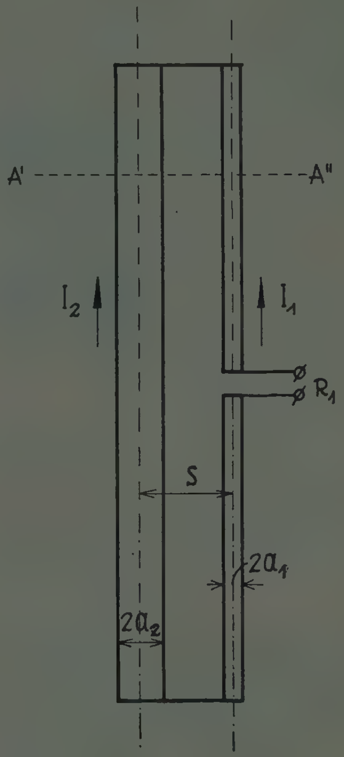

Fig. 1—The folded dipole.

As shown in reference 1 the impedance transformation of the folded dipole is given by:

$$\frac{R_1}{R_0} = (1+n)^2, \qquad (1)$$

where R_1 is the input impedance of the folded dipole, R_0 the input impedance of the simple dipole, and n stands for the current ratio:

$$n = \frac{I_2}{I_1}. \qquad (2)$$

From the equation of continuity ($\operatorname{Div} j = -i\omega\rho$)

$$\frac{j_2}{j_1} = \frac{\rho_2}{\rho_1}, \qquad (3)$$

or:

$$\frac{I_2}{I_1} = \frac{q_2}{q_1}. \qquad (4)$$

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* Original manuscript received by the Institute April 27, 1953.

[1] R. Guertler, "Impedance transformation in folded dipoles," Proc. I.R.E., Vol. 38, p. 1042; September, 1950.

The impedance transformation is thus defined by the ratio of charges in the elements. The ratio of charges q_1 and q_2 can be found with the aid of the condition that the scalar potential on the conductor boundary remains constant at any cross section of the dipole (A′A″—Fig. 1).

Assuming a sinusoidal charge distribution along the Z direction (Fig. 2).

$$q(x'y'z') = f(x'y') \cos \frac{2\pi z'}{\lambda}, \qquad (5)$$

where q represents the charge distribution and $x'y'z'$—a point of the surface of the conductors. The scalar potential is given by:

$$\phi = \frac{1}{4\pi\epsilon_0} \int\int\int \frac{1}{r} [q] dx'dy'dz', \qquad (6)$$

the integration should be carried out over the whole surface of the conductors; $[q]$ is the retarded charge, and $r = \sqrt{(x-x')^2+(y-y')^2+(z-z')^2}$. By a proper choice of the time origin, the retarded charge can be written as:

$$[q] = f(x'y') \cos \frac{2\pi z'}{\lambda} \cos \frac{2\pi \gamma_1}{\lambda}, \qquad (7)$$

substituting (7) into (6) and considering the plane $z=0$:

$$\phi_{z=0} = \frac{1}{4\pi\epsilon_0} \int\int f(x'y') \cdot \left\{ \int_0^{\lambda/2} \frac{\cos \frac{2\pi r_1}{\lambda} \cos \frac{2\gamma\pi z'}{\lambda}}{\gamma_1} dz' \right\} dx'dy' \qquad (8)$$

where:

$$r_1 = \sqrt{(x-x')^2 + (y-y')^2 + z'^2}.$$

The integral in parenthesis has been solved by Guertler,[1] who has shown that when $\sqrt{(x-x')^2+(y-y')^2} \ll \lambda/2$ the following approximation holds:

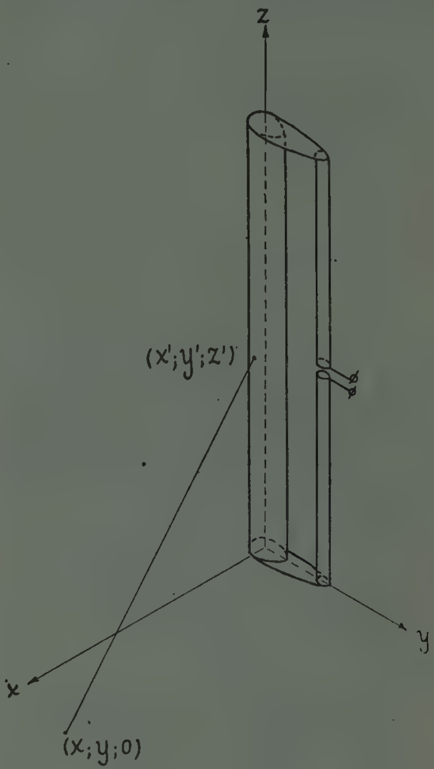

Fig. 2—Folded dipole with arbitrary geometry of cross section.

$$\int_0^{\lambda/2} \frac{\cos \frac{2\pi z'}{\lambda} \cos \frac{2\pi r_1}{\lambda}}{r_1} dz' \doteq \ln \frac{A}{[(x-x')^2 + (y-y')^2]^{1/2}}, \qquad (9)$$

where A is a constant; substituting (9) into (8) we have:

$$\phi_{z=0} = \frac{1}{4\pi\epsilon} \int\int f(x'y') \cdot \ln \frac{A}{[(x-x')^2 + (y-y')^2]^{1/2}} dx'dy' \qquad (10)$$

but this is the well known equation of the two dimentional electrostatic field.[2]

From (10) together with the boundary condition that ϕ should be constant on the surface of the elements at any cross section, we may conclude: When the cross section dimensions of the folded dipole are small (in comparison with $\lambda/2$) the ratio of charges in the elements is approximately equal to the electrostatic charge ratio of infinitely long conductors having the same cross section as the folded dipole, all the conductors being on the same potential.

Considering the elements as quasi-line charges[3] immediately leads to the solution given in reference 1. The charge ratio for arbitrary cross sections can be found conveniently with the aid of the electrolytic tank.

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[2] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., p. 219; 1941.

[3] E. Weber, "Electromagnetic Fields," John Wiley and Sons, vol. 1, p. 106; 1950.

Contributors to Proceedings of the I.R.E.

M. Burgett, Jr.

Monte I. Burgett, Jr. (S'42–A'43–SM'51) was born in Midland, Ark. on March 8, 1921. He attended Hendrix College and Georgia School of Technology where in 1943 he received the BS in EE degree, Co-operative Plan. He received his MS in EE degree from the University of Pennsylvania in 1951.

In 1943 Mr. Burgett joined Philco Corporation and has been with that company ever since. During the war years, he worked on airborne radar and bombing equipment, and afterward on auto radio and television tuner development. Since 1950 he has been working on color systems and receivers in the Research Department where he serves as a project engineer.

Mr. Burgett joined the faculty of Drexel Institute of Technology Evening College in 1948, and organized the electronics option of Electrical Engineering. From 1950 to 1952, he served as Assistant Head of the Drexel Electrical Engineering Department.

For a photograph and biography of David K. Cheng, see page 998 of the August, 1952 issue of the Proceedings of the I.R.E.

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For a photograph and biography of Marvin Chodorow, see page 163 of the January, 1953 issue of the Proceedings of the I.R.E.

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G. C. Dacey

G. C. Dacey was born on January 23, 1921, in Chicago, Illinois. He received the B.S. degree in electrical engineering from University of Illinois in 1942. Upon graduation he joined the staff of the Westinghouse Research Laboratories where he worked on the development of "Resnatrons" and high-power magnetrons.

At the end of the war he enrolled at the California Institute of Technology and completed his studies for the Ph.D. degree in 1950. He remained at the Institute for a year as the A. O. Smith Post-Doctoral Fellow studying radio tracer methods in the investigation of metals.

In 1951, Dr. Dacey joined the staff of the Bell Telephone Laboratories as a member of the transistor physics group.

Dr. Dacey is a member of A.P.S., Eta Kappa Nu, Tau Beta Pi, Sigma Tau, Phi KappaPhi, and Sigma Xi.

❖

F. H. Dickson

Frederic H. Dickson (S'38–A'43) was born in Bristol, Colorado, on January 10, 1916. He received the B.S. in electrical engineering from Oregon State College in 1939, where he was a graduate assistant in electrical engineering until 1940. Prior to entry in military service in 1942, he was an instructor in the Air Corps Technical Schools. Mr. Dickson served as an assistant radio officer, Allied Forces Headquarters and as Assistant Signal Officer, 15th Army Group in the European theatre and returned from overseas as Executive Officer and later Commanding Officer of the Signal Corps Radio Propagation Unit. Upon release from military service in 1946, he joined the Office of the Chief Signal Officer and has been active in application engineering and operational research in the field of radio-wave propagation. He is presently Chief of the Radio Propagation Section.

Mr. Dickson is a member of the USA National Committee, URSI, Chairman of US Commission IV, URSI and is a registered professional engineer in the District of Columbia.

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For a photograph and biography of JOHN J. EGLI, see page 163 of the January, 1953 issue of the PNOCEEDINGS OF THE I.R.E.

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R. A. GALBRAITH

Ralph A. Galbraith was born on August 1, 1911 in Ash Grove, Mo. He received the B.S. degree in 1933 from the University of Missouri and the Ph.D. in Electrical Engineering from Yale University in 1937.

Dr. Galbraith has held the posts of research engineer for the Detroit Edison Company in 1937, and research editor, the following year. From 1938 until 1939, he was an instructor in electrical engineering at the University of Missouri, and from 1939 until 1944, an assistant professor to associate professor of electrical engineering at the University of Texas. During the summer of 1941 he was research engineer at Century Electric Company. He was a staff member at M.I.T. Radar School from 1944 until 1946, and professor of electrical engineering at Georgia Institute of Technology in 1946. For the next five years Dr. Galbraith was chairman in the department of electrical engineering at Syracuse University.

From that time until now he has been dean of the college of engineering at Syracuse University.

❖

E. L. GINZTON

Edward L. Ginzton (S'39–A'40–SM'46–F'51) was born in Russia on December 27, 1915, and came to the United States in 1929. He received the B.S. degree in electrical engineering from the University of California in 1936, and the M.S. degree from the same institution in 1937. For graduate study at Stanford University, he received the E.E. degree in 1938, and the Ph.D. degree in 1940.

From 1937 until 1939 Dr. Ginzton acted as assistant in teaching and research at Stanford University, and in 1940 he became a research associate in the physics department.

From 1940 until 1946 Dr. Ginzton was employed in the research laboratories of the Sperry Gyroscope Company. From 1946 to 1948, he has been an assistant professor of applied physics at Stanford University, and an Assistant Director of the Microwave Laboratory. In 1948 he became an associate professor, and in 1951, professor of applied physics and electrical engineering.

Dr. Ginzton has been the Director of the Microwave Laboratory since 1949. He has also been a member of the Executive Committee of the Board of Directors of Varian Associates from 1948 to date. He is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

❖

W. J. GRUEN

Wolf J. Gruen (M'47) was born in Berlin, Germany, on September 21, 1914. He received the Dipl. Ing. E.T.H. degree in electrical engineering from the Swiss Federal Ins. of Technology in Zurich, in 1940, and the M.E.E. degree from Syracuse University in 1952.

From 1944 to 1946 Mr. Gruen was in charge of the Laboratory of the Long Lines Department of the Ericsson Telephone Company in Mexico City, Mexico. In 1946 he immigrated to this country and joined the Advanced FM Development Group of the Hazeltine Electronics Corporation in New York.

Since August 1947, Mr. Gruen has been associated with the Receiver Department of the General Electric Company. He is currently a Section Engineer in charge of a development section, engaged in monochrome and color-television circuit development.

Mr. Gruen is a member of Panel 15, NTSC, and has been active on several subcommittees of the NTSC. He currently serves on the Membership Committee of the Syracuse Section of the I.R.E. He is a member of the Scientific Research Society of America, RESA.

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For a photograph and biography of J. W. HERBSTREIT, see page 606 of the May, 1952 issue of the PROCEEDINGS OF THE I.R.E.

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W. R. HEWLETT

William R. Hewlett (S'35–A'38–SM'47–F'48) was born in 1913 at Ann Arbor, Mich. He received the A.B. degree from Stanford University in 1934, and the M.S. degree from the Massachusetts Institute of Technology in 1936. In 1939 he received the E.E. degree from Stanford University, after spending the period from 1936 to 1938 engaged in electromedical research in Palo Alto, Calif.

In 1939, Mr. Hewlett joined David Packard in starting the Hewlett-Packard Company in Palo Alto. During the war he was on active duty with the Army, first assigned to the office of the Chief Signal Officer, and then to the New Developments Division of the War Department's Special Staff in Washington, D.C. Since 1946, he has been associated with the Hewlett-Packard Company.

Mr. Hewlett is a member of Sigma Xi and the American Institute of Electrical Engineers. He received the I.R.E. Fellow award "for his initiative in the development of special radio measuring techniques."

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ROLAND HOLT

Roland B. Holt was born on July 21, 1920, in San Antonio, Texas. He received his A.B. degree from the University of Texas in 1940; his M.A. degree from the University of Texas in 1942; and his Ph.D. degree from Harvard University in 1947.

During the period 1942–1950, Dr. Holt held the positions of research associate, National Research Council Predoctoral Fellow, Instructor and Assistant Professor of Physics at Harvard University.

In 1950 Dr. Holt became Director of the Cyclotron Laboratory at Harvard University, a position he held until 1952. During the fall of 1951 he became President of Scientific Specialties Corporation, the position he currently holds as well as President of Transistor Products, Inc.

Ph.D. thesis work involved a study of the role of hydrogen peroxide in the thermal reaction between hydrogen and oxygen.

During World War II at the Radio Research Laboratory of Harvard University, Dr. Holt did work in the field of radio countermeasures.

Dr. Holt is a member of Phi Beta Kappa, Sigma Xi, I.R.E. (senior member), and the A.P.S.

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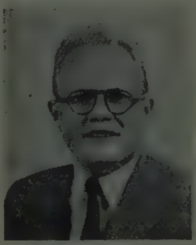
R. A. JOHNSON

Richard A. Johnson was born in Worcester, Massachusetts on June 30, 1926. For two years during the war he was in Navy Radar program after which he received the A.B. degree in physics from Dartmouth in 1948. He continued his studies at Harvard University, receiving the M.S. Degree in 1950 and the Ph.D. Degree in Engineering Science and Applied Physics in 1952. At Harvard he did research in gas discharge physics and the theory of random processes. At present, Dr. Johnson is affiliated with Transistor Products, Inc., He is a member of A.P.S., Phi Beta Kappa and Sigma Xi.

Josef Kates

Josef Kates (S'45–A'52) was born on May 5, 1921, in Vienna, Austria. He received the B.A. degree in the mathematics and physics honor course in 1948, the M.A. degree in applied mathematics in 1949, and the Ph.D. degree in physics in 1951, all at the University of Toronto.

From 1944 to 1948, Dr. Kates worked on problems related to vacuum tubes with Rogers Electronic Tubes, Ltd., Toronto. Since 1948 he has been engaged in the development and design of digital computers with the Computation Centre of the University of Toronto.

Dr. Kates is a member of the Association of Professional Engineers of Ontario, the Canadian Association of Physicists, and the Canadian Mathematical Congress.

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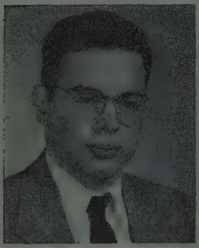

Leonard H. King

Leonard H. King (S'46–A'49) was born in New York City on July 27, 1925. After serving in the Army from October 1943 to May 1946, he entered the Massachusetts Institute of Technology and received a B.S. degree in electrical engineering in June 1950. Mr. King became a research assistant in the M.I.T. Instrumentation Laboratory and received the M.S. degree in electrical engineering in June 1951. Since then, he has worked as a research engineer in the airborne fire control section of the M.I.T. Instrumentation Laboratory and presently is in charge of the control aspects of a prototype fire control system.

Mr. King is an associate member of Sigma Xi.

❖

Geoffrey Knight

Geoffrey Knight, Jr. was born on July 15, 1922, in Washington D. C. He received his A.B. degree, magna cum laude, from Amherst College in 1943 and his Ph.D. in theoretical physics from M.I.T. in 1949.

From 1949 to 1951 Dr. Knight worked in the fields of transistor technology and transistor applications for International Business Machines Corporation. In 1951 he joined Scientific Specialties Corporation, eventually transferring to its affiliate, Transistor Products, Inc. Here he has been chiefly concerned with the design of equipment for measuring transistor characteristics and with transistor circuit applications.

He is a member of the A.P.S., Phi Beta Kappa, and Sigma Xi.

❖

E. J. Nalos

Ervin Joseph Nalos was born in Prague, Czechoslovakia, September 10, 1924. He graduated from King Edward High School, Vancouver, Canada, in 1941. In 1946 he received the degree of Bachelor of Applied Science in electrical engineering, and in 1947, the degree of Master of Applied Science from the University of British Columbia. He continued graduate study as a research and teaching assistant at the Microwave Laboratory at Stanford University under a Sperry Gyroscope Company fellowship, obtaining his Ph.D. in 1951.

Since then Dr. Nalos has been a Research Associate at the Microwave Laboratory, where he is active in high power microwave tube research and the student laboratory instruction program. He is a member of Sigma Xi.

❖

R. J. Nordlund

Raymond J. Nordlund (M'46–SM'50) was born in New Effington, South Dakota, on April 20, 1911. He received the B.S. in EE Degree from South Dakota State College in 1933, and completed basic requirements for the M.S. Degree in Radiation at Ohio State University in 1948.

From March, 1934, until June, 1941, Mr. Nordlund was engaged in various phases of highway design and engineering for the South Dakota State Highway Commission. In June, 1941, he joined the staff of the Signal Corps Aircraft Radio Laboratory, Wright Field, Ohio, undertaking research in "night effect" in radio-compass airborne navigation.

During 1942 Mr. Nordlund was engaged in component development for the first radar-navigation systems to be used by the U. S. Air Force, SCR-521, SCR-517, and AN/APQ-5. From March, 1943, to January, 1946, he served as project engineer for development of a fire-control radar system and a high-resolution radar bombing system in Radar Laboratory, Wright Field. From January, 1946, to October, 1948, he served as Chief, Ground Position Indicating Systems Unit and as Asst. Chief, Bombing Branch, Aircraft Radiation Laboratory. He was Chief, Strategic Bombing Branch, Armament Laboratory from October, 1948, to April, 1951. At the present time his post is that of Bombing Consultant, Armament Laboratory, Wright Air Development Center, Dayton, Ohio.

Mr. Nordlund is a member of Sigma Tau, and has served on a sub-panel of the Research and Development Board for several years. He holds the rank of Major, USAFR.

❖

Ian M. Ross

Ian M. Ross was born in Southport, England, on August 15, 1927. He received the B.A. and M.A. degrees in engineering from the University of Cambridge in 1948 and 1952. He received the Ph.D. degree from Cambridge in 1952 for an investigation of low-frequency noise in vacuum tubes and semiconductors.

Dr. Ross joined the staff of Bell Telephone Laboratories in February 1952 and has been since that time a member of the transistor physics group.

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For a photograph and biography of Eugene A. Sands, see page 1254 of the October, 1952 issue of the Proceedings of the I.R.E.

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Norman R. Scott

Norman R. Scott (SM'46) was born in Brooklyn, N. Y., on May 15, 1918. He received B.S. and M.S. degrees in 1941 from Massachusetts Institute of Technology, where he was enrolled in the cooperative course in Electrical Enginering. He worked at the General Electric Co. during his undergraduate training and for a short time after graduation. In 1941 he entered the U. S. Army, serving for five years. He concluded his service with the rank of Major and the position of Chief of Special Projects Laboratory. He joined the faculty of the Dept. of Electrical Engineering at the University of Illinois in 1946 and received his Ph.D. there in 1950. He is now Asst. Professor of Electrical Engineering at the University of Michigan.

Dr. Scott is a member of Sigma Xi, Eta Kappa Nu, and Phi Kappa Phi.

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L. S. Sheingold

Leonard S. Sheingold (S'46–A'50) was born in Boston, Mass., on January 14, 1921. He served in the United States Army from 1942 to 1945.

In 1947 Mr. Sheingold received the B.E.E. degree from Syracuse University. In 1948 and 1949, he was an instructor in electrical engineering at Syracause University. He received the M.E.E. degree from Syracuse University in 1949. From 1949 to 1952, he was a graduate student in the Dept.

of Engineering Science and Applied Physics at Harvard University.

Mr. Sheingold obtained the A.M. degree from Harvard in 1950 and the Ph.D. degree in Applied Physics in 1953.

During the summers of 1950 and 1951, Dr. Sheingold was employed as a Microwave Engineer for Sylvania Electrical Products Inc., and served as a consultant during 1952. At present Dr. Sheingold is employed as an engineering specialist with the Engineering Lab. of Sylvania Electric Products Inc., Boston, Mass.

Dr. Sheingold is a member of Sigma Xi, Sigma Pi Sigma, and is a member of the papers committee of the Boston Section of the I.R.E.

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For a photograph and biography of Jacob Shekel, see page 1611 of the November, 1952, issue of the Proceedings of the I.R.E.

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Blanchard D. Smith, Jr. (S'45–A'50) was born in New Orleans, La. on August 22, 1925. He received the B.S. degree in electrical engineering from the Georgia Institute of Technology in 1945 and the M.S. degree in electrical engineering from the Massachusetts Institute of Technology in 1948.

B. D. Smith, Jr.

Mr. Smith was a commissioned officer in the U. S. Navy as a radar instructor. He was a research assistant at the Research Laboratory of Electronics, M.I.T., working on telemetering systems and pulse-code modulation.

Since 1948 Mr. Smith has been empolyed by Melpar, Inc. At present he is a consulting project engineer on electronic research and development projects.

Mr. Smith is a member of Phi Kappa Phi, Tau Beta Pi, Eta Kappa Nu, Sigma Xi.

◆

Ernest K. Smith, Jr. (A'46–M'48) was born in Peiping, China, on May 31, 1922. He came to the United States in 1940 and entered Swarthmore College, where he received the B.A. in Physics in 1944. After graduation, he went into the Army and spent the following two years with the Signal Corps Radio Propagation Unit, Baltimore, Maryland. Mr. Smith joined the Engineering Department of the Mutual Broadcasting System in 1946 and later became Chief of the Plans and Allocations Division. In 1949, he entered the School of Electrical Engineering at Cornell University and received the M.S. degree in 1951.

Ernest K. Smith,

Mr. Smith was employed by the Central Radio Propagation Laboratory of the National Bureau of Standards in Boulder, Colorado from June 1951 to September 1952. He is now a candidate for the Ph.D. degree at Cornell University.

◆

Harry Sohon (A'28–VA'39–SM'50) was born on November 24, 1904, in New York City. He received the E.E. degree from Cornell University in 1926. After graduation he worked for the General Electric Company for two years.

Harry Sohon

Mr. Sohon returned to Cornell on a McMullin Fellowship and received the M.E.E. degree in 1930. Joining the staff at Cornell as Instructor, he earned the Ph.D. degree in 1932. He continued teaching at Cornell until 1942 when he was appointed Asst. Professor of Electrical Engineering at the University of Connecticut where he remained a year before accepting a similar appointment at the Moore School, University of Pennsylvania, from 1943 to 1945.

Following a year as advanced development engineer in the aviation engineering section of RCA-Victor, Mr. Sohon returned to the Moore School in 1946. In 1951 he was appointed Associate Professor of Electrical Engineering and also Associate Supervisor of Research at the Moore School.

Mr. Sohon is licensed as a professional engineer in the State of Pennsylvania and is a member of Eta Kappa Nu, Phi Kappa Phi, Sigma Xi, the Engineers Club of Philadelphia, the American Society for Engineering Education, the American Association of University Professors, and the American Institute of Electrical Engineers.

◆

Seymour Stein (S'53) was born in Brooklyn, New York, on April 4, 1928. He received the B.E.E. degree from City College, New York, in 1949, and the M.S. degree in Applied Science from Harvard University in 1950. He is now studying at Harvard toward a Ph.D. in in Applied Science.

S. Stein

Mr. Stein was a Teaching Fellow at Harvard during 1950–51, and is now an RCA Fellow in Electronics, under the National Research Council (1951–1953).

Mr. Stein is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

◆

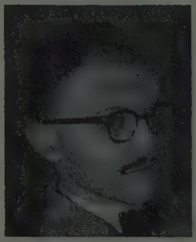
James E. Storer was born in Buffalo, N. Y., on October 26, 1927. He received the A.B. degree in physics in June, 1947, from Cornell University, followed by the M.S. and Ph.D. degrees in 1948 and 1951, respectively, both from the Dept. of Engineering Sciences and Applied Physics at Harvard University. While at Harvard, he held an Atomic Energy Commission Fellowship. He is a member of the Sigma Xi Society.

James E. Storer

Dr. Storer was Research Fellow at the Electronics Research Lab., 1951–53, and Lecturer in the Div. of Applied Science, Harvard, 1952–53. Currently he is Asst. Professor in the Div. of Applied Science, doing research in the fields of Electromagnetic Theory and Random Processes.

◆

Stanley Weintraub was born in New York, N. Y., on April 4, 1922. He entered the City College of New York in 1939, leaving in 1942 to join the Signal Corps Radar Laboratory, enlisting in the Signal Corps the same year. He served as an airborne radar technician in the India-Burma theater from 1943 to 1946. Mr. Weintraub returned to CCNY, receiving the B.S. degree in Physics in 1947. From 1947 until 1949 he was in the Mechanical Instruments Section of the National Bureau of Standards, Boulder, Colo. Since 1949 he has been with the Central Radio Propagation Laboratory of the Bureau where he is now a member of the radio-meteorological group of the Tropospheric Propagation Research Section.


S. Weintraub

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For a photograph and biography of G. S. Wickizer, see page 1731 of the December, 1952 issue of the Proceedings of the I.R.E.

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For a photograph and biography of Lotfi A. Zadeh, see page 1128 of the September, 1952, issue of the Proceedings of the I.R.E.

Institute News and Radio Notes

FOREIGN MEMBERS MAY PAY DUES WITH UNESCO COUPONS

The attention of foreign members is called to the availability of UNESCO Coupons for paying IRE membership dues. This coupon is issued by the United Nations Educational, Scientific and Cultural Organization to facilitate payments of this nature between UNESCO member countries.

UNESCO Coupons may be purchased with local currency from the national Distributing Body of each participating country or from UNESCO regional offices. The addresses of Distributing Bodies and UNESCO offices are given below.

Distributing Bodies

Austria—Aussenhandelstelle fur Buch und Graphik, Grunangergasse 4, Vienna I; **British Colonial and Trust Territories**—Unesco Book Coupons, c/o Book Tokens Ltd., 28 Little Russell Street, London, W.C.1; **Burma**—The Secretary, National Commission Secretariat, Government of Burma, 545–547 Merchant Street, Rangoon; **Cambodia**—Ministere de l'Education Nationale et de la Jeunesse, Phnom Penh; **Ceylon**—The Secretary, National Commission for Unesco, Ministry of Education, Colombo; **Czechoslovakia**—Orbis Company, 37 Narodni, Prague 1; **Egypt**—Administration of General Culture, Ministry of Education, Cairo; **France**—Direction des Bibliotheques de France, 55 rue Saint Dominique, Paris-7^{e}.

French Colonial Territories—Direction des Bibliotheques de France, 55 rue Saint-Dominique, Paris-7^{e}; **Germany**—(Bundesrepublik) Notgemeinschaft der deutschen Wissenschaft, Buechelstrasse 55, Bad Godesberg bei Bonn; **Hungary**—Kultura, P.O.B. 149, Budapest 62; **India**—Ministry of Education, New Delhi 3; **Indonesia**—Ministry of Education and Culture, Djalan Tijlatjap 4, Djakarta; **Israel**—Dr. G. J. Ehrlich, Import Licensing Office, Ministry of Education and Culture, Hakirya; **Italy**—Commissione Nazionale dell'Unesco, Villa Massimo, Via di Villa Massimo, Rome; **Japan**—Society for the Promotion of Science, c/o Science Council of Japan, Ueno Park, Tokyo, Japan; **Pakistan**—The Minister of Education and Industries (Education Division), Government of Pakistan, Karachi.

Persia—Persian National Commission for Unesco, Avenue du Nusee, Teheran; **Syria**—The Secretary, Syrian National Commission, Ministry of Education, Damascus; **Thailand**—The Thailand National Commission for Unesco, Ministry of Education, Bangkok; **Union of South Africa**—The Secretary, Department of Education, Arts and Science, New Standard Bank Buildings, Pretoria; **U.S.A.**—Unesco Office, UN Building, Room 2201, New York 17, N. Y.; **United Kingdom**—Unesco Book Coupons, c/o Book Tokens Ltd., 28 Little Russell Street, London, W.C.1; **Yugoslavia**—Secretaire do la Commission Nationale de la Republique populaire federative de Yougoslavie pour l'Unesco Moskovska 51, Belgrade.

Calendar of
COMING EVENTS

- **IRE-ONR-Univ. of Calif. Symposium on Statistical Methods in Communication Engineering, Berkeley, Calif., August 17–18**
- **IRE Western Convention and Electronic Show, Civic Auditorium, San Francisco, Calif., August 19–21**
- International Sight and Sound Exposition and Audio Fair, Chicago, Ill., September 1–3
- Conference on Nuclear Engineering, University of California, Berkeley, September 9–11
- **National Electronics Conference, Hotel Sherman, Chicago, Ill., September 28–30**
- **URSI Fall Technical Meeting, Ottawa, Canada, October 5–8**
- **PGUE-Acoustical Society Symposium, Case Institute of Technology, Cleveland, Ohio, October 15–17**
- **1953 IRE-RTMA Radio Fall Meeting, Toronto, Ont., October 26–28**
- Conference on Radio Meteorology, University of Texas, Austin, November 9–12
- **Joint IRE-AIEE 6th Annual Conference on Electronic Instrumentation in Medicine and Nucleonics, New York City, November 18–20**

UNESCO Offices

(Covering Afghanistan, Burma, Ceylon, China, Egypt, Hashemite, Jordan, India, Indochina, Indonesia, Iraq, Israel, Japan, Korea, Lebanon, Persia, Philippines, Pakistan, Saudi Arabia, Syria, Thailand, Turkey.)

East Asia—Unesco Science Co-operation Office, United Nations Building, 106 Whangpoo Road, Shanghai, China and Unesco Science Co-operation Office, United Nations Building, Padre Faura, Manila, Philippines; **Middle East**—Unesco Science Co-operation Office, 8 Sh. el Salamlik, Garden City, Cairo, Egypt and Unesco Science Co-operation Office, Istanbul Teknik Universitesi, Gumussuyu, Istanbul, Turkey; **South Asia**—Unesco Sciency Co-operation Office, University Buildings, Delhi, India; **South-East Asia**—Unesco Science Co-operation Office, Merdeka Seletan 11 Pav., Djakarta, Indonesia.

IRE SECTIONS FORMED IN LONG ISLAND AND ELMIRA

At its May 5 meeting the IRE Board of Directors approved the establishment of the Long Island Section (formerly a Subsection of New York Section) in Region 2 and the Elmira-Corning Section in Region 4.

These raise IRE Sections to 67.

IRE AUTHORS FIRST CALL

Prospective authors for the National Convention please submit: name, address, title of paper, 100-word abstract and informational up to 500 words (both in triplicate), to B. R. Lester, IRE, 1 East 79 St., New York, N. Y. by November.

W. R. G. BAKER RECIPIENT OF TWO AWARDS

Dr. W. R. G. Baker (A'19–F'28), General Electric Company vice president and past president of the IRE, was recently presented the Medal of Freedom by the Honorable Earl D. Johnson, Under Secretary of the Army, during ceremonies in the Pentagon, and the Medal of Honor by the Radio-Television Manufacturers Association at their annual convention in Chicago.

Dr. Baker was awarded the Medal of Freedom for accelerating the application of electronics to the solution of Army research and development problems. He led a mission of leading scientists and industrialists to Korea in the summer of 1952, to study the problem of utilizing electronic devices and principles to the maximum extent, thereby increasing the effectiveness of the individual soldier and reducing the cost of human life.

ULTRASONICS ENGINEERING

The First Administrative Committee meeting of the Professional Group on Ultrasonics Engineering was held on May 6, 1953 in Washington, D. C.

Members of the committee are: Chairman of the Group: A. L. Lane, Naval Ordnance Laboratory; Vice-chairman, M. D. Fagen, Bell Telephone Laboratories; Secretary, J. Hunter, John Carroll University; and W. G. Cady, California Institute of Technology; W. J. Fry, University of Illinois; Frank Massa, Massa Laboratories; Oskar Mattiat, Clevite-Brush Development Co., P. L. Smith, Naval Research Lab.; W. J. Mayo-Wells, Johns Hopkins Univ.

Committee chairmen are as follows: Paper Procurement and Publications, O. Mattiat; Membership, M. D. Fagen; Sectional Activities, J. Hunter; Finance, P. L. Smith; Legal, W. J. Mayo-Wells.

Plans were made to sponsor a session on ultrasonics engineering at the National Electronics Conference in September.

Institute News and Radio Notes

NATIONAL TELEMETERING CONFERENCE BANQUET

Among those who attended the National Telemetering Conference jointly sponsored by the IRE, AIEE, IAS and ISA were (l. to r.) R. S. Gardner of AIEE; Dr. Harvard Hull of IAS, conference treasurer; D. J. E. Hobson, Stanford Research Institute, banquet speaker; Kipling Adams, General Radio Co.; John Doremus, Allen G. Caldwell Manufacturing Co., conference chairman; W. A. Wildhack of ISA; L. G. Cumming, IRE Technical Secretary.

Organization for Engineering Unity

The following report to the membership is published at the direction of the IRE Board of Directors.—*The Editor*

For several years, earnest and detailed efforts have been made by a group of engineers broadly representative of leading engineering societies to promote unity in the engineering profession. Their further purpose was to afford both a forum and a spokesman, so to speak, for engineers on matters of such broad scope as might transcend the defined fields of individual engineering societies.

The body chosen in mid-1949 to study these matters, under the general auspices of the Engineers Joint Council and about fifteen collaborating engineering societies (which had appointed their representatives), was known as the "Exploratory Group to Consider the Increased Unity of the Engineering Profession." The IRE was represented on this Exploratory Group.

It was evident at once that any concrete procedures for an increase in the "unity" of the engineering profession were necessarily complex, multi-element matters.

More specifically, three major difficulties were immediately encountered. The first of these may be briefly described as follows. The attractiveness of the broad concept of engineering unity was diminished as soon as an attempt was made to define at least some of the *specific* objectives and procedures of the unity organization. Indeed, up to this time there are some persons who, while sympathetic to the broad concept of engineering unification, find no definite or concrete proposals or aims which seem sufficient at this time to command support. The second difficulty arose from the conflict between a unification based on engineering-society membership and control and a consolidation based on individual membership and control, independent of the existing societies. The third problem sprang from varying definitions of the scope and authority of individual engineering societies. Some of these regard themselves as restricted closely to scientific and technical matters. Others wish to go further afield, in some cases impinging on the fields of political, legislative, or even labor-representational activities. Some societies do not regard themselves as authorized to speak for their membership on any but highly technical matters, except perhaps after a special validating vote by the membership. Others take almost the opposite stand as to their right to commit their membership on a wide variety of subjects.

These problems were brought into sharp focus for the IRE by the receipt of an invitation from the Engineers Joint Council (EJC) to join a broadened and modified version of that body planned to function as an engineering unity organization.

One of the objectives of the expanded EJC was the development of sound public policies respecting national and international affairs. This aim could be interpreted as permitting the direct influencing of legislation—a course of action believed to be favored by some in the EJC. Further, a simple majority vote of the Council might be regarded as sufficient to initiate such action. The IRE has consistently refrained, as a matter of established policy, from any such action either as a society or through its Sections.

The new body would also administer, on behalf of the engineering profession, those activities authorized by a majority of the constituent societies of the Council. The implication would seem to be that while the individual engineer is to be represented, only the constituent societies would enjoy individual membership. The IRE has consistently held to the position that the IRE is not entitled to represent its members as individual members of the engineering profession. Presumably an IRE constitutional amendment would be needed to authorize the IRE Board of Directors thus to represent its membership more widely and to extend that power of representation to another body of its choice.

While the IRE does not take a stand as a society even on technical matters when dealing with outside agencies, it does provide facilities for independent bodies such as the Joint Technical Advisory Committee. This method of operation has proved satisfactory (and is similar to the excellent precedent of the Royal Society, Great Britain, which, throughout its long and distinguished history, has carefully avoided ever taking a position as a society, while assisting wherever possible in the formation and work of any commission established to deal with national problems).

At the March 27, 1953 meeting of the IRE Board of Directors, the Board considered at length the EJC invitation and expressed its viewpoint in the following unanimously approved resolution:

Whereas, Our study of the Constitution of the Engineers Joint Council, adopted December 7, 1952, has revealed that acceptance by The Institute of Radio Engineers of the invitation to join the Engineers Joint Council would necessitate abrogation of two basic policies of long standing: one, that the Institute shall not engage directly or indirectly in legislative activity; and the other, that the Institute, as a society, is not entitled to represent its members as individual members of the engineering profession or as a body on social or general professional matters, and shall not, as a society, take a stand even on technical matters; therefore be it

Resolved, That, after careful deliberation, the Board of Directors has come to the conclusion that it is not desirable to reverse these policies at this time in order to permit affiliation with the Engineers Joint Council; and be it further

Resolved, That, therefore, The Institute of Radio Engineers must regretfully decline the invitation to join the Engineers Joint Council.

However, the Board looked favorably upon co-operative activity with the EJC on any specific problems of mutual interest in which the Institute can engage within the above stated framework. The Board further emphasized that the action taken was believed by it to be in the best interests of the Institute and its members. It should be added that, in the opinion of the Board, a unity organization of the type envisioned could function more appropriately and effectively as an individual membership society rather than as a coalition of professional engineering societies. And, further, the Board directed that the preceding information should appear in the PROCEEDINGS OF THE I.R.E. as a report to the membership

Professional Group News

Airborne Electronics

The Philadelphia Chapter of the Professional Group on Airborne Electronics elected W. W. Felton, Franklin Institute, president; R. J. Esser, Raymond Rosen Engineering Products, vice-president; and C. E. Dolberg, Philco Corp., secretary. At a recent meeting at the Naval Air Development Center in Johnsville, Pa., C. S. Vasaka, engineer with the Center, gave a paper entitled the "Effects of Radio Interference on Airborne Electronics." He discussed the necessity for reduction of radio interference, the methods of reduction, testing, and the test equipment involved.

Antennas and Propagation

The new officers and administrative committee of the Professional Group on Antennas and Propagation are: Chairman, P. S. Carter; Vice Chairman, D. C. Ports; Secretary-Treasurer, P. H. Smith; George Sinclair; A. H. Waynick; J. B. Smyth; H. G. Booker; H. A. Finke; J. S. Brown; L. J. Chu; R. B. Jacques; L. C. Van Atta; H. W. Wells.

Audio

The Cincinnati Chapter of the Professional Group on Audio met under the Chairmanship of R. E. Kolo recently at the Hotel Gibson, and Engineering Society Hdqts. G. J. Schulmerich, president of Schulmerich Electronics Inc. spoke about "Carillons." He discussed the characteristics of the tones produced by cast bells and covered the methods used in his carillonic bells to secure comparable results. C. A. Maynard, vice president, Indiana Steel Products Co., talked about "Magnetic Structures for Audio and Acoustic Devices." He covered the characteristics of, and magnetic and physical limitations of, permanent magnet materials used in structures for audio and acoustic devices, and the design and application of magnets to such structures.

The newly formed Houston Chapter held an organizational meeting in May at which L. A. Geddes was elected head. D. P. Carlton demonstrated his two latest versions of small folded horns, one with an 8-inch speaker and a tweeter, and other with two 8-inch speakers. (Drawings are available upon request, care of Humble Oil.) He also demonstrated a console with a McIntosh amplifier, Rek-O-Nut turntable, and switching facilities permitting the selection of various output impedances and particular speaker. The PGA tapescript and slides of "A Single Ended Push-Pull Amplifier" by Peterson and Sinclair of General Radio were shown, and Roy Brougher demonstrated a modified version of this type of amplifier. James Hallenburg showed a message repeater. The next meeting will coincide with the beginning of binaural FM broadcasts, probably in August.

The Philadelphia Chapter recently held two meetings in the Edison Building presided over by S. C. Spielman, and W. G. Chaney. Lowell Good, RCA, spoke on "Design Factors for Audio Amplifiers," presenting what performance characteristics are needed for a suitable amplifier and how these requirements are being met in commercial amplifiers. He covered power, distortion, noise, stability, reliability, and efficiency as they affect satisfactory design. H. H. Scott, president of H. H. Scott, Inc., gave a paper "Component Integration of Sound Reproducing Systems" in which he discussed the over-all problems of high-fidelity audio systems, particularly the contributions of the transducer, amplifier, and the loudspeaker system. He covered the application of dynamic noise suppression to phonographic reproduction from the new low noise records in detail.

Broadcast Transmission Systems

The Boston Chapter of the Professional Group on Broadcast Transmission Systems met recently at the studios of WCOP with P. K. Baldwin as Chairman. D. W. Bodle of Bell Telephone Labs spoke on "Lightning Protection of Transmitter Plants." Staff members of 15 broadcast stations attended. New officers of the Chapter are Sidney Stadig, WBZ-TV, Chairman; Hollis Gray, WHDH, Vice-Chairman; and Fearing Pratt, N.E. Tel. & Tel., Secretary.

The Chicago Chapter met recently under the Chairmanship of Stephen Bushman; F. M. Dukat, Receiving Tube Division of Raytheon Mfg. Co., was the speaker. In "Some Crystal Application Problems in Radio and Television Receivers," he discussed the use of crystals as low frequency multipliers and uhf mixers, using slides of graphs as illustration. He brought out that video detector performance is affected by the back voltage characteristic of the crystal, the temperature at which the crystal is operating, and the capacity and resistance loading of the crystal upon the detector circuit. Multiplier crystals are similar to video detector crystals, but are chosen for their higher front-to-back resistance ratio and better high frequency characteristics. UHF mixer crystals have a much lower front resistance, lower back voltage range and less capacity than the detector type crystals. Sufficient oscillator current must be injected into the mixer crystal to achieve optimum conversion efficiency. However, the crystal cannot be driven too far in the reverse current direction because of the increase in noise temperature of the crystal. Officers for the coming year are: Chairman, J. T. Williams; Vice-Chairman, K. R. Decho; Secretary, J. F. White.

The Los Angeles Chapter met recently in the Conference Room of the IAS Building with D. E. Foster presiding. A. A. Barco, RCA Labs, spoke on "Developments in the Application of Transistors." At a later meeting E. P. Thias and Robert Eland of Standard Coil Products talked about "An 82-Channel Turret Tuner." Officers for the coming year will be: Chairman, Stanley Cutler, Pacific Mercury TV Mfg. Co.; Vice-Chairman, J. T. McAllister, Hoffman Radio Corp.; and Secretary-Treasurer, A. J. Arman, Stevens Co.

Circuit Theory

The Chicago Chapter of the Professional Group on Circuit Theory met at the Western Society of Engineers Building recently under the Chairmanship of L. E. Pepperberg. Speaker of the evening was B. S. Parmet, project engineer with Motorola, Inc., who discussed "The Design of Maximally Flat Amplitude and Time Delay Filters."

Electron Devices
North Shore Chapter Formed

The establishment of a North Shore Chapter of the Professional Group on Electron Devices (Boston Section) has been approved by the Executive Committee of the Institute.

Interested members please contact G. E. Carter, Jr., Bomac Labs., Inc., Salem Road, Beverly, Mass.

New Group on Electronic Component Parts

The need for a professional group concerned with the field of component parts has caused such a group to be established. Officially called the "Electronic Component Parts Professional Group," its interest will be confined to the characteristics, limitations, applications, development, performance, and reliability of component parts. Membership will be open to all IRE members. Engineers interested are invited to join.

Officers of the Group are: Chairman, F. A. Paul, supervisor, reliability group, Northrup Aircraft Co., Hawthorne, Calif.; Vice-Chairman, A. W. Rogers, chief, components and materials branch, Squier Signal Lab., SCEL, Fort Monmouth, N. J.; Secretary-Treasurer, C. G. Walance, Research & Devel. Labs., Hughes Aircraft Co., Culver City, Calif. Administrative committee members are: M. J. Ainsworth, Bendix Corp., 7250 Laurel Canyon Blvd., N. Hollywood, Calif.; J. T. Brothers, components section, Philco Corp., Philadelphia 34, Penn.; J. A. Csepely, Air Arm Div., Westinghouse Electric Corp., Friendship International Airport, Baltimore, Md.; F. B. Haynes, supervisor Electronics Product Service Sect., Glenn L. Martin Co., Baltimore; L. K. Lee, Advanced Techniques Lab., Stanford Research Institute, Stanford, Calif.; H. E. May, tube application co-ordinator, Motorola, Inc., 4545 Augusta Blvd., Chicago 51, Ill.; L. Podolsky, technical asst. to the president, Sprague Electric Co., No. Adams, Mass.; G. Shapiro, Electronics Div., National Bureau of Standards, Washington 25, D. C.; W. G. Tuller, Melpar Electronics, Inc., 452 Swann Ave., Alexandria, Va.

Engineering Management

The Chicago Chapter of the Professional Group on Engineering Management met recently with F. W. Schor as Chairman. A. W. Graf, patent lawyer, spoke about "Patent Law as it Concerns Employers and Employees." A panel of ten members discussed the subject further with Mr. Graf following his talk.

TECHNICAL COMMITTEE NOTES

Under the chairmanship of D. C. Ports the **Antennas and Waveguides** Committee met on May 20. There was a discussion of Standards Committee actions on Waveguide terms. The latter part of the meeting was taken up with a discussion of Component definitions.

The **Circuits** Committee convened May 22 under the chairmanship of W. R. Bennett. W. A. Lynch, Chairman of Subcommittee 4.7 reported recommendations made at a meeting of his subcommittee concerning the list of feedback terms Appendix "A" 53 IRE 4. M3. The committee reviewed the above definitions.

The **Electron Devices** Committee convened on May 8 under the chairmanship of G. D. O'Neill. In the absence of R. B. Janes there was no report on Phototube Methods of Test. If the revised phototube definitions are satisfactory to Mr. Janes they will be started on Grand Tour. A report was relayed from Mr. Janes that his subcommittee has considered the latest revised proposals (Storage Tube Definitions) and that nearly all replies are in from a questionnaire sent to interested parties outside the subcommittee.

The **Facsimile** Committee met on May 8 under its new Chairman, Henry Burkhard. The activities of the committee under its former Chairman were reviewed and the following summary was given by J. H. Hackenberg. A general revision of the "1942 Standards on Facsimile: Definitions of Terms" has brought up to date the definitions, which now encompass electronic as well as mechanical scanning techniques. Included are new recording techniques which separate the recording field from the closely related art of television. A list was given of the terms that had been acted upon, revised, and approved in present form by the committee. Also listed were the definitions under discussion but not acted upon. Another item under consideration by the committee is a test chart. On March 6, 1953 it was agreed that essentials of a test chart include four indicators of amplitude or density fidelity, and fifteen which serve to indicate the fidelity of dimension or positioning of the picture elements. A readability chart has been under discussion by the committee and some studies have been made but no definite action taken. The most recent proposal described in the Minutes of March 6, 1953 meeting looks promising, and may supplant the readability chart. After reading the summary of activities the scope of the committee was reviewed.

The **Information Theory and Modulation Systems** Committee met on April 21 under the chairmanship of W. G. Tuller. It was moved, seconded, and passed that the definitions (under consideration by the committee), in agreement with the broader point of view, are on communication theory rather than information theory and should be so headed. The remainder of the meeting was spent in a discussion of the above definitions.

The **Navigation Aids** Committee convened on May 25 under the chairmanship of P. C. Sandretto. Some time was spent discussing the definitions tentatively approved at the April 27 meeting. Revisions were made in a number of them. Final review of the proposed definitions has now been completed.

On May 11 the **Video Techniques** Committee met under the chairmanship of W. J. Poch. The proposed standard on the Methods of Measurement of Aspect Ratio and Geometric Distortion tentatively approved at the last meeting was reviewed. After several minor corrections it was given final approval for consideration by the Standards Committee by unanimous vote of those present. A. J. Baracket reported that it was the intention of his subcommittee to continue its consideration of Methods of Geometric Distortion with emphasis on effects due to changes in scanning-beam velocity. It is understood that the Subcommittee on Video Signal Transmission is preparing a report on a method of measurement involving the use of a step signal in combination with a super-imposed high-frequency sinewave. The Chairman reported that L. G. Grignon's paper entitled "Television Video Recording on Film" was returned to the Subcommittee on Utilization with the suggestion that minor revisions be made to reflect the changes in the art since the paper was prepared. The proposed Standards on Methods of Measurement of Pulse Quantities, Part I previously distributed to the committee members was discussed. Definitions approved at the previous meeting were again reviewed. The list of those definitions in the Video Techniques field previously issued in published form were reviewed and approved with a few exceptions.

Western Electronic Show & Convention*

CIVIC AUDITORIUM, SAN FRANCISCO, CALIF.—AUGUST 19–21, 1953

Wednesday, 10:00 A.M., August 19

SESSION I: ELECTRON DEVICES I

Chairman, M. Chodorow, Stanford Univ.

"A 1.8–4 KMC High Gain Wide-Band TWT Amplifier," S. F. Kaisel, L. A. Roberts, and R. P. Lagerstrom, Stanford Univ.

"A Wide-Band Power Mixer Tube," H. R. Johnson, Hughes Research and Development Laboratories

"A Wide-Tuning Range Microwave Oscillator/Amplifier," J. L. Putz and W. R. Luebke, Stanford Univ.

"Helix-Type Backward-Wave Oscillators," D. A. Watkins, Stanford Univ.

"Cross-Modulation in Traveling-Wave Amplifiers," A. W. C. Nation and J. W. Christie, Univ. of Washington

SESSION II: COMPUTERS I

Chairman, L. D. Stevens, IBM

"A Series-to-Parallel Data Converter," G. A. Neff, R. L. Sink, and H. E. Burke, Consolidated Engineering Corp.

"A New Analog-to-Digital Voltage Converter," J. Zweizig, California Institute of Technology

"An Analog-to-Digital Conversion System with Printed Decimal Read-Out," J. L. Lindesmith, Clary Multiplier Corp.

"An Analog-to-Digital Converter," A. D. Scarbrough, Hughes Aircraft Co.

"The Analyzing Reader," D. H. Shepard, Intelligent Machines Research Corp.

SESSION III: NOISE AND SIGNAL SPECTRA

Chairman, W. W. Harman, Stanford Univ.

"Instantaneous or Measurable Frequency Spectra," A. D. Watt and V. J. Zurick, National Bureau of Standards

"The Response of Linear Systems to Non-Gaussian Noise," B. Gold and G. O. Young, Hughes Research and Development Laboratories

"Linear Detection of Nonstationary Noise-Like Signals," Ralph Deutsch, Hughes Research & Development Laboratories

"A System of Noise Analysis," S. D. Wanlass and D. M. Jacob, Hughes Research and Development Laboratories

Wednesday, 2:30 P.M., August 19

SESSION IV: COMPUTERS II

Chairman, Torben Meisling, Univ. of California

"An Improved Reading System for Magnetically Recorded Digital Data," Samuel Lubkin, Underwood Corp.

"Magnetic Materials for Digital Computers," D. R. Brown, Massachusetts Institute of Technology

"Panel Discussion on the Relative Merits of Different Memory Types," *Moderator:* P. L. Morton, Univ. of California

SESSION V: AIRBORNE ELECTRONICS

Chairman, A. R. Ellis, Stanford Research Institute

"The Air Navigation Development Board's Program for the Development of the Common System of Air Navigation and Traffic Control," D. K. Martin, Air Navigation Development Board

"The Measurement of Performance of Airborne Voice-Modulated Communication Systems," E. J. Moore and John Taylor, Stanford Research Institute

"Corona Interference Reduction by Polarity Discrimination," M. M. Newman, Lightning and Transients Research Institute

"Magnetic Amplifiers and Their Applications," Victor Boros and David Seddman, Polytechnic Research and Development Co.

"Airborne Weather Radar for Transport

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* Sponsored jointly by IRE Region 7 and the West Coast Electronic Manufacturers Association.

Aircraft," Richard White, Trans World Airlines

Session VI: Instrumentation I

Chairman, D. B. Sinclair, General Radio Co.

"The Application of Counter Techniques to Precision Frequency Measurements," A. F. Boff, Berkeley Scientific Div., Beckman Instruments

"Two Timing-Circuit Innovations," H. B. Brooks, Hughes Aircraft Co.

"Strain Gage Oscillator," E. A. Varallo, Raymond Rosen Engineering Products

"Measurements of Time Jitter in Trains of Video Pulses," J. L. Fitch and R. R. Buss, Stanford Univ.

"A Peak-Reading Vacuum-Tube Voltmeter Which Has a Long Decay Time and is Capable of Measuring the Amplitude of Short Pulses," L. S. Cutler, Gertsch Products

Session VII: Electron Devices

Chairman, T. Moreno, Varian Associates

"Convection Current Noise—Theory and Experiment," S. V. Yadavalli, Univ. of California

"Microwave Oscillator Stability," George Hetland and R. R. Buss, Stanford Univ.

"Air-Coolers for High-Power Vacuum Tubes," A. L. London, Stanford Univ.

"A High-Power High-Gain X-Band Pulsed Amplifier Klystron and Driver Multiplier," L. Zitelli, Varian Associates

"A High-Gain K-Band Amplifier," W. G. Abraham and F. L. Salisbury, Varian Associates

"Operating Behavior of High-Power Pulsed Klystrons," John Jasberg, Stanford Univ.

Thursday, 10:00 A.M., August 20

Session VIII: Transistors

Chairman, (to be announced)

"Recovery-Time Measurements on Point-Contact Germanium Diodes," Morgan McMahon, T. E. Firle, and J. F. Roach, Hughes Aircraft Co.

"A Point-Emitter Junction-Collector Transistor," R. H. Kingston, MIT

"Measurement of the Small Signal Parameters of Transistors," Geoffrey Knight, Jr., R. A. Johnson, and R. B. Holt, Transistor Products, Inc.

"Rapid Determination of Some Electrical Properties of Semiconductors," Luther Davis, Jr., Lawrence Rubin, and W. D. Straub, Raytheon Manufacturing Co.

Session IX: Antennas I

Chairman, A. S. Dunbar, Dalmo Victor Co.

"Design and Performance of Rotationally Symmetric Feeds for Paraboloidal Reflectors," H. W. Haas, R. W. Dressel, and R. D. Ewing, New Mexico College of Agriculture and Mechanic Arts

"A New Antenna Feed Having Equal *E* and *H* Plane Patterns," Alvin Chlavin, Hughes Aircraft Co.

"Waveguide Slot Arrays of Large Squint Angle," R. J. Adams and A. M. Lide, Naval Research Laboratory

"The Impedance Properties of Narrow Radiating Slots in the Broad Face of Rectangular Waveguides," Arthur A. Oliner, Polytechnic Institute of Brooklyn

"Principles of Spiral Scanners for Equal Pulse Distribution," J. Richard Huynen, Dalmo Victor Co.

"Boresight Theory for Homogeneous Dielectric Radomes," M. C. Horton, W. E. L. Boyce, and E. O. Hartig, Goodyear Aircraft Corp.

Thursday, 10:00 A.M., August 20

Session X: Nuclear Radiation Measurements

Chairman, H. S. Bright, U. S. Naval Radiological Defense Laboratory

(Tentative Topics)

"Gamma and Electron Spectrometry with Crystals at High Energy"

"A Discussion of Some Unsolved Instrumentation Problems in Nuclear Physics"

"The Current Status of Radiation Detector Development"

"Neutron Source Standardization"

Session XI: Servomechanisms

Chairman, O. J. Smith, Univ. of California

"Nonlinear Control Systems with Random Inputs," R. C. Booton, Jr., MIT

"Comparison of Linear and Nonlinear Servomechanism Response," T. M. Stout, Univ. of Washington

"Time Quantization in a Feedback System," J. F. Waddell and H. D. Morris, Univ. of California

"Stability of Feedback Systems Using a Dual Locus Diagram," Paul Jones, California Institute of Technology

"Geometrical Interpretation of the Response of Linear Systems to Special Inputs," J. R. Moore, North American Aviation

Thursday, 2:30 P.M., August 20

Session XII: Transistor Circuits

Chairman, H. M. Zeidler, Stanford Research Institute

"Recent Developments in Transistors," Irving Wolff, RCA

"Transistor Shift Registers," R. H. Baker, I. L. Lebow, and R. E. McMahon, MIT,

"A Point Contact Transistor VHF FM Transmitter," D. E. Thomas, Bell Telephone Laboratories

"A Four-Digit Transistor Accumulator," D. J. Eckl, MIT

"A Transistor Feedback Amplifier for Carrier Frequency Applications," J. C. Lozier and D. D. Cherry, Bell Telephone Laboratories

Session XIII: Microwave Theory & Techniques

Chairman, E. T. Jaynes, Stanford Univ.

"Mode Representations in Open and Closed Uniform Waveguides," Nathan Marcuvitz, Polytechnic Institute of Brooklyn

"Applications of Coupled Helices," P. D. Lacy, Hewlett-Packard Co.

"New Applications of Faraday Rotation in Waveguides," A. G. Fox, M. T. Weiss, and S. E. Miller, Bell Telephone Laboratories

"Nonreciprocal Circuits Comprising Ferrite-Loaded Rectangular Waveguides," A. G. Fox, M. T. Weiss, and S. E. Miller, Bell Telephone Laboratories

"The Generation of Electromagnetic Oscillations in the Microwave Region Using an Adiabatic Kind of Amplification," Gedalia Held, Univ. of California

Session XIV: Antennas II

Chairman, J. T. Bolljahn, Stanford Research Institute

"Arrays of Closely Spaced Nonresonant Slots," R. J. Stegen and R. H. Reed, Hughes Aircraft Co.

"Diffraction Theory and the Patterns of Suppressed Antennas," George Sinclair, Univ. of Toronto

"Beam-Shaping and Optimum Bandwidth Methods Applied to UHF TV Transmitting Antennas," John Ruze and J. E. Martin, The Gabriel Laboratories

"Voltage Protection of Isolated Cap Aircraft Antennas," R. L. Tanner, Stanford Research Institute

"A Slotted Cylinder Omnirange Projector," J. P. Shanklin, Collins Radio Co.

Session XV: Servomechanism Equipment

Chairman, O. J. Smith, U. of Cal.

"A Method of Analysis and Synthesis of Closed-Loop Servo Systems Containing Small Discontinuous Type Nonlinearities," D. T. McRuer and R. G. Halliday

"Optimum Lead-Controller Synthesis in Feedback-Control Systems, L. G. Walters, U. of Cal., Los Angeles

"Predictor Servomechanisms," Lawrence Silva, U. of Cal., Berkeley

"A Complex-Frequency Generator, Yale Barkan, Hughes Aircraft Corp.

"Development of Precision Servo-Controlled Potentiometer-Setting Equipment," L. L. Gordon, Electronic Associated, Inc.

Thursday, 8:00 P.M., August 20

Session XVI: Color Television

Chairman, W. H. Doherty, Bell Telephone Laboratories

Speakers:

W. R. G. Baker, Vice-President, General Electric Co. and Chairman of the National Television System Committee

Donald G. Fink, Director of Research, Philco Corp., and Chairman, Panel 12 of the NTSC

Friday, 10:00 A.M., August 21

Session XVII: Audio Symposium

Chairman, Vincent Salmon, Stanford Research Institute

Panel:

"Microphones," W. B. Snow, Western Electro-Acoustic Laboratory

"Recording," F. G. Lennert, Ampex Corp.

"Amplifiers," A. N. Curtis, RCA Victor Division

"Loudspeakers," R. H. Smith, Univ. of Cal.

Session XVIII: Circuit Theory I

Chairman, B. J. Bennett, Stanford Research Institute

"The Practical Implication and Applications of Formal Network Theory," D. F. Tuttle, Stanford Univ.

"Design of a Simple Band-Pass Amplifier with Approximate Ideal Frequency Characteristics," W. E. Bradley, Philco Corp.

"Quasi-Distortionless Filter Functions," J. L. Stewart, Univ. of Michigan

"Fluctuation Noise Theory as Applied to Circuit Design," T. S. George, Air Force Missile Test Center, Patrick Air Force Base

SESSION XIX: MICROWAVE THEORY & TECHNIQUES II

Chairman, J. R. Whinnery, Univ. of California

"A Microwave Oscillograph," R. C. Honey, Stanford Research Institute

"Instrumentation of Microwave Electron Resonance in Magnetic Fields," R. C. Mackey and W. D. Hershberger, Univ. California

"An Improved Cross Guide Directional Coupler," H. J. Riblet, Microwave Development Laboratories

"Two Novel Types of Waveguide Switches," Amasa Pratt, Century Electronics, Division of Century Metalcraft Corp.

"Broad-Banding Circular Polarizing Transducers," D. L. Margerum, Microwave Engineering Co.

SESSION XX: PROPAGATION—GENERAL

Chairman, A. M. Peterson, Stanford Univ.

"Waveguiding on Surfaces with and without Loss," F. J. Zucker, Air Force Cambridge Research Center

"A New Solution to the Ionospheric Wave Equation," A. J. Mallinckrodt, The Ralph M. Parsons Co.

"Ionosphere Sounding by Cross-Correlation Techniques," P. B. Gallagher and A. M. Peterson, Stanford Univ.

"The Long-Distance Horizontal Directivity of a 13.7 MC Antenna," Richard Silberstein, National Bureau of Standards

Friday, 2:30 P.M., August 21

SESSION XXI: PROPAGATION, VHF & UHF

Chairman, J. B. Smyth, U. S. Naval Electronics Laboratory

"Results of Tropospheric Propagation Measurements on Frequencies from 92 to 1046 MC at the Cheyenne Mountain Field Station," A. F. Barghausen and K. O. Hornberg, National Bureau of Standards

"Characteristic of an 8.6 mm Radio Transmission Path," C. W. Tolbert and A. W. Straiton, The University of Texas

"An Investigation of the Variation of VHF Field Strength Beyond Line-of-Sight," G. H. Keitel and H. M. Swarm, Univ. of Washington

"Air-to-Air Propagation—Experimental and Theoretical Results," M. S. Wong, Wright Air Development Center

"The Role of Angular Distance in Tropospheric Radio-Wave Propagation," K. A. Norton, National Bureau of Standards

"Normal Propagation of Short Radio Waves Well Beyond the Horizon," T. J. Carroll and R. M. Ring, MIT

SESSION XXII: CIRCUIT THEORY II

Chairman, G. L. Matthaei, Univ. of California

"Solving Physical Systems with Very Large Number of Variables in Easy Stages," Gabriel Kron, General Electric Co.

"Matrix Analysis of Linear Time-Varying Circuits," L. A. Pipes, Univ. of California and U. S. Naval Ordnance Test Station

"Unbalanced RLC Networks Containing Only One Resistance and One Real Transformer," Louis Weinberg, Hughes Research and Development Labs.

"An Iterative Method for Network Synthesis," R. E. Scott, MIT and R. L. Blanchard, Transonics, Inc.

SESSION XXIII: INSTRUMENTATION II

Chairman, W. B. Wholey, Hewlett-Packard Co.

"Measurement Problems in VHF-UHF Television Antenna Systems," R. A. Soderman, General Radio Co.

"An Auto Impedance Meter for VHF-UHF," John Ebert, Polytechnic Research Development Co.

"A Ratiometer," N. L. Pappas, Hewlett-Packard Co.

"An Improved Method of Measuring the Current Amplification of Junction-Type Transistors," F. R. Stansel, Bell Telephone Laboratories

SESSION XXIV: AUDIO

Chairman, Roy Long, Stanford Research Institute

"Stereophonic Tape System," R. H. Snyder, Ampex Electric Corp.

"Application and Suggestions for Research Concerning Acoustical Problems in Medical Areas," J. K. Hilliard, Altec Lansing Corp.

"An Investigation of the Air Chamber of Horn-Type Loudspeakers," Bob H. Smith, Univ. of California

"A Simple Calibration Technique for Low Sensitivity Transducers," W. J. Galloway, Signal Corps Engineering Laboratories and Univ. of California

Friday, 8:00 P.M., August 21

SESSION XXV: MEDICAL ELECTRONICS

Chairman, A. J. Morris, USN Office of Naval Research

"Area Display by Electronic Mapping, Especially of the Electrical Activity of the Heart," Stanford Goldman, Syracuse Univ.

"Electronic Mapping of the Brain," A. R. Tunturi, Univ. of Oregon Medical School

"Radioactive Tracer Mapping," H. O. Anger and C. A. Tobias, Donner Laboratory of Medical Physics and the Univ. of California

IRE People

Isaac L. Auerbach (S'46–M'49–SM'51) has been appointed manager of the Special Electronic Equipment Department of the Burroughs Adding Machine Co. He will primarily be concerned with defense equipment and magnetic development research for the government.

I. L. AUERBACH

Formerly a research engineer, Mr. Auerbach has been engaged in the design of electronic data processing systems and initiated work on magnetic components and circuits. He was responsible for the design of the static magnetic memory system for the ENIAC.

A native of Philadelphia, Mr. Auerbach received the B.S. degree in electrical engineering from Drexel Institute of Technology in 1943. He did graduate work at the Massachusetts Institute of Technology and Harvard University, from which he received the M.S. in applied physics in 1947.

From 1943 to 1946 Mr. Auerbach served as a Navy radar officer, in 1945 assigned to the Naval Research Laboratory for work on IF systems. Prior to joining Burroughs he worked on the BINAC and UNIVAC systems at the Eckert-Mauchly Division of Remington Rand, Inc., and was a tube engineer with the Electronic Tube Corporation.

Mr. Auerbach is chairman of the Philadelphia chapter of the Professional Group on Electronic Computers. He is also a member of Eta Kappa Nu, the American Institute of Electrical Engineers, Association for Computing Machinery, the Scientific Research Society of America, and a registered professional engineer in the state of Pennsylvania.

❖

Robert B. Beetham (S'34–A'36) has been named works manager of the Hicksville, N. Y. plant of Amperex Electronic Corp.

Mr. Beetham had been associated with the Airborne Instrument Laboratory, Mineola, N. Y., since 1949, where he held the positions of executive assistant to the vice president in charge of research and engineering, and since 1951 executive assistant director of research and engineering.

R. B. BEETHAM

Mr. Beetham was born in Jewett, Ohio on July 6, 1904. He received the B.S. degree in electrical engineering from Ohio State University in 1934 and the M.S. in 1936.

Following graduation Mr. Beetham became affiliated with the Collins Radio Company, Ames, Ia. A radio engineer, he eventually became assistant to the president.

Mr. Beetham has served on several IRE committees and is an associate member of the A.I.E. and a member of Eta Kappa Nu.

IRE People

Gano Dunn (F'15–L'50) internationally known engineer and builder and president of the J. G. White Engineering Corporation of New York City, died recently.

Gano Dunn

Mr. Dunn was born in New York City on October 18, 1870. He received the B.S. degree from the College of the City of New York in 1889, and from Columbia University in 1891 the first degree of the first class in electrical engineering awarded degrees in the United States. He later attended RCA Institutes, and held a first-class commercial radio operator's license.

Following graduation, Mr. Dunn joined the Crocker-Wheeler Company and assisted in the design of motors, generators, transformers and other electrical apparatus. He subsequently became chief engineer, vice-president and a director of that company.

Mr. Dunn became associated with the J. G. White Engineering Corporation in 1911 as vice-president in charge of engineering and construction. As president from 1913 on, he directed many outstanding construction projects, among which were 13 transatlantic radio stations, the original Government aviation station at Langley Field, and roads, bridges, power plants, and irrigation systems in Latin America.

Mr. Dunn was elected a director of the Radio Corporation of America in 1938. He was also a director of RCA Communications, Inc., and the National Broadcasting Company. Active in the furtherance of education, he was chairman of the trustees and former president of Cooper Union for the Advancement of Science and Art. He was a member of the Visiting Committee of Harvard Engineering School and former president of the American Institute of Electrical Engineers.

Mr. Dunn was an officer or delegate on many scientific committees and special governmental commissions, including the Science Advisory Board of President Roosevelt. President Hoover offered the directorship of the Bureau of Standards to him. He was asked to consider the secretaryship of the Smithsonian Institution, and twice declined the presidency of the Massachusetts Institute of Technology.

Among the awards presented to Mr. Dunn were the Townsend Harris Medal of the College of the City of New York, the Thomas A. Edison Medal of the American Institute of Electrical Engineers, the Egleston Medal of Columbia University, Hoover Medal of the National Engineering Societies, Modern Pioneer Award of the National Association of Manufacturers, and the Peter Cooper Medal of the Cooper Union.

Honorary degrees of Mr. Dunn include the D.Sc. from Columbia, Rutgers, Lehigh, and New York Universities, the College of the City of New York, and the LL.D. degree from Bowdoin College, Brunswick, Maine.

Mr. Dunn was honorary secretary for the United States of the Institute of Electrical Engineers and honorary president of the Pan-American Society.

Mr. Dunn was a fellow of the Royal Microscopy Society and the New York Academy of Sciences, and a member of the American Society of Civil Engineers, American Society of Mechanical Engineers, American Academy of Arts and Sciences, American Philosophical Society, Optical Society of America, New York Historical Society, and the New York Zoological Society, Phi Beta Kappa, Sigma Xi, Phi Beta Pi, and Delta Kappa Epsilon.

❖

George F. Maedel (M'43–SM'43), formerly vice president and general superintendent of the RCA Institutes, has been elevated to the presidency of the Institutes.

George F. Maedel

Mr. G. F. Maedel joined RCA Institutes in 1933 as the first instructor of the mathematics department. He transferred to the radio frequency department in 1936, where he became chief instructor in 1940. Mr. Maedel was appointed assistant superintendent of the Institutes in 1944, and became superintendent in 1947. The next year he was elected to his former position as vice president and general superintendent.

A native of Brooklyn, Mr. Maedel was born there on July 21, 1903. He received the A.B. degree in 1924 and the E.E. degree in 1926 from Columbia University.

Mr. Maedel began his career with the New York Edison Company, where he worked in their planning bureau for a year following graduation. He joined the New York Telephone Company in 1927, to do traffic engineering. From 1931–1933 he headed the Audio Products Engineering Company, where he was concerned with installing public address systems.

Mr. Maedel is a licensed professional engineer in the state of New York.

Paul N. Bergquist (M'48), broadcast consultant, has been appointed field sales representative for broadcast equipment for the RCA Victor Division of the Radio Corporation of America.

Mr. Bergquist was born in Spokane, Wash., on October 8, 1919. He received the B.S. degree in electrical engineering from State College, Pullman, Wash., in 1942. During his college years he was a radio operator for various northwestern radio stations.

In 1942 he joined the U. S. Navy as a radar officer and attended radar school at Harvard University and Massachusetts Institute of Technology. He served as a training and maintenance officer and later as staff radar officer of the Naval Air Advanced Training Command responsible for the training in and maintenance of radar of nineteen Naval Air stations.

After separating from the service in 1946, Mr. Bergquist became a radio engineer with Glen D. Gillett and Associates, later a partner of Gillett and Bergquist.

❖

Major William S. Dawson (A'49) has assumed command of the 1300th Student Squadron of the Air Resupply and Communications Service, Mountain Home Air Force Base, Idaho.

Major Dawson recently returned from Japan, where he was Director of Plans and Requirements for 1808th Airways and Air Communications Service Wing. He was responsible for supervision of planning and engineering of communications facilities and electronic navigational aids throughout the Pacific and Korea. He had been Deputy Director until that position was created.

During the past sixteen years Major Dawson has served in military aviation in meteorological and communications assignments such as commanding officer of an AACS Squadron and at the Headquarters of the former Army Air Corps.

Major Dawson is a member of the American Geophysical Union, American Institute of Physics, American Meteorological Society, American Association for Advancement of Science, and the Institute of Aeronautical Sciences.

❖

J. Walton Colvin (M'51) has been appointed manager of government sales for the Communication Division of Bendix Radio.

Mr. Colvin joined Bendix in 1942 as a vhf instructor. He successively became technical advisor to the Mediterranean Headquarters of the Air Force from 1944–46, aviation service engineer, and aviation sales engineer. Since 1948 he has been merchandising manager, dealing in Air Force sales.

A native of Somerset, Va., Mr. Colvin was born on December 30, 1914. He is a graduate of the National Radio Institute of Washington, D. C., and Northwestern Institute, Ames, Iowa.

Previous to his association with Bendix Mr. Colvin operated his own radio, electrical sales, and service business.

IRE People

David E. Cavenaugh (A'46), Aircraft Transformer Corporation, Long Branch, N. J., has been elevated from the position of assistant to chief engineer.

Mr. Cavenaugh was born on September 26, 1919 in North Carolina. He received the B.S. degree in electrical engineering from the Georgia School of Technology, Atlanta, Ga., in 1940.

Following graduation Mr. Cavenaugh worked briefly for the Atlantic Coast Line Railroad Company as a special apprentice before joining the U. S. Army Signal Corps. As a captain Mr. Cavenaugh was concerned with communication of all types and the organization, planning, and training for a signal operation battalion until 1944.

Mr. Cavenaugh joined Bell Telephone Laboratories in 1945. As a member of the power apparatus group he was concerned with small power transformer design and high voltage cathode-ray tube power supplies. He joined the Aircraft Transformer Corporation in 1952.

Mr. Cavenaugh is a member of the AIEE.

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F. W. Albertson

Fred W. Albertson (A'33–M'44–SM'49), radio engineer, attorney at law, and partner of Dow, Lohnes and Albertson, Washington, D.C., has been elected president of the Federal Communications Bar Association.

Mr. Albertson was born in Fairgrove, Mich., on September 29, 1908. He received the A.B. degree from the University of Michigan in 1931 and the LL.B. in 1934. He was admitted to the bar in Michigan the same year, the District of Columbia in 1935, the Supreme Court of the United States in 1940, U. S. Treasury Department in 1935, Federal Communications Commission, Interstate Commerce Commission, and Securities and Exchange Commission in 1936, and other boards and commissions.

Mr. Albertson has been an amateur radio operator since 1920 and holder of a commercial license since 1925, from which date he did work as chief operator and engineer until 1934. He became a registered professional electrical-communications engineer in the District of Columbia in 1952.

Since 1934 Mr. Albertson has been radio law counsel for numerous radio, telegraph, telephone and broadcast companies and stations, consulting radio engineering firms, and radio manufacturing companies. These include Westinghouse Electric Co., Radio Engineering Laboratories, Inc., Gulf Research and Development Co., Major Edwin H. Armstrong, FM Development Foundation, Paul F. Godley Co., Raymond M. Wilmotte, and others.

Mr. Albertson has been active in IRE activities of the Detroit and Washington Sections and nationally, having served in various offices of the Washington Section and on the national committees of Public Relations, Awards, and Constitution and By-Laws. He is currently on the Board of Editors and Membership committees.

Mr. Albertson is a member and former president of the Washington Radio Club and the University of Michigan Radio Club (of which he was co-founder). He is a charter member and has held various offices in the Engineers Club of Washington and Federal Communications Bar Association.

Among the other professional societies to which Mr. Albertson belongs are the Academy of Arts and Sciences, the Philosophical Society of Washington, the Radio Pioneers, Delta Theta Pi, American Bar Association, Bar Association of the District of Columbia, and the American Judicature Society.

◆

D. W. Gunn

D. W. Gunn (M'47) has been appointed to the newly created post of assistant general sales manager of the radio and television tube division of Sylvania Electric Products, Inc. He will continue his duties as sales equipment manager.

Mr. Gunn has been with Sylvania since 1931, when he started as a factory engineer in Salem, Mass. In 1936 he became a sales engineer in Chicago, and since then has worked successively as a quality engineer, supervisor of quality control, eastern district sales engineer, east central district sales engineer, and assistant to the general sales manager. He became equipment sales manager in 1951.

Born in New York state in 1907, Mr. Gunn received the B.E.E. degree from Northeastern University, Boston, Mass. in 1929. He then joined the New England Power Co., later the Raytheon Manufacturing Co., Newton, Mass. as a production engineer.

◆

Lawrence A. Hyland (A'29), vice president of Bendix Aviation Corp., has been placed in charge of its engineering program.

Formerly in charge of Bendix research, Mr. Hyland has been a vice president since 1949. He joined Bendix in 1935, when the Radio Research Company which he founded became affiliated with Bendix, and in 1937 became general manager of radio operations, and vice president of Bendix Radio Corp. In 1943 he was elevated to executive engineer of the Aviation Corp.

Mr. Hyland was born in Woodlawn, Nova Scotia on August 26, 1897. In 1917 he joined the U. S. Army, serving as a radio sergeant. In 1920 he transferred to the aviation branch of the U. S. Navy as a chief radioman. He participated in the first blind landing of a flying boat by radio in 1921. He was given a special discharge in 1926 to become an assistant, later associate radio engineer at the Naval Research Laboratory. Here he first observed and proved the principle of radar detection of aircraft in 1931.

In 1950 Mr. Hyland received the Navy's highest civilian honor, the Distinguished Public Service Award, for his "great service to science and to the welfare of the United States through his early contribution to the development of radar." He is credited with more than 40 inventions, including the radio-shielded spark plug which made possible modern aircraft communications by clearing up interference, the Navy radio wing-loop direction finder, and frequency indicators.

Mr. Hyland is a member of the Guided Missiles Committee of the U. S. Research and Development Board, the Industry and Advisory Board of the U. S. Air Force Arnold Engineering Development Center, the Institute of Naval Engineers, and the Society of Automotive Engineers.

◆

J. K. Mitchell

John K. Mitchell (A'40–SM'46) has joined the newly organized Barth Engineering and Manufacturing Co. of Milldale, Conn. as vice president and general sales manager. This company was called the Barth Manufacturing Co. under former ownership.

Mr. Mitchell was formerly associated with Sperry Gyroscope Co., Great Neck, L. I. Since 1946 he had been a sales representative, and more recently that company's representative to the U. S. Navy.

Mr. Mitchell was born in East Orange, N. J., on November 10, 1906. He received the B.S. degree in electrical engineering from the Newark College of Engineering in 1936. While serving in the Navy from 1942–45 he attended Bowdoin College's pre-radar course, the Massachusetts Institute of Technology's Radar School and the Naval Research Laboratory's special project school.

Mr. Mitchell then became an anti-jamming officer in the Bureau of Aeronautics, conducted field tests of anti-jamming devices, was an aide to a Navy Liaison officer at the Radio Research Laboratory and Radiation Laboratory in Cambridge, Mass., and finally officer in charge of the Radio Research Navy Office of the Radio Research Laboratory at Harvard University.

Before joining the Navy in 1942 Mr. Mitchell had served from 1923–27 as a radioman in the submarine service. He then worked for a year with the New York Telephone Co. before joining the New Jersey Telephone Co. Until 1938 he held various positions with this company, at which time he became an instructor in electrical engineering at the Newark Technical School.

Mr. Mitchell is a military member of the Society of Naval Engineers.

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

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The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

ACOUSTICS AND AUDIO FREQUENCIES

534.2:534.011 **1874**

An Acoustic Gyrator—W. E. Kock. (*Arch. elekt. Übertragung*, vol. 7, p. 106; Feb. 1953. In English.) A device is described which is the acoustic analogue of the microwave gyrator described by Hogan (1233 of 1952). The required nonreciprocal rotation of the plane of polarization of transverse acoustic waves propagated in a tube is accomplished by rotating a section of the tube at high speed.

534.21-13 **1875**

On Sound Waves of Finite Amplitude—E. T. Copson. (*Proc. Roy. Soc. A.*, vol. 216, pp. 539–547; Feb. 24, 1953.) An analytical solution of Riemann's equations for the one-dimensional propagation of sound waves of finite amplitude in a gas obeying the adiabatic law $p=k\rho\gamma$ is obtained for any value of the parameter γ.

534.231 **1876**

The Sound Field of a Piston Source—E. W. Guptill (*Canad. Jour. Phys.*, vol. 31, pp. 393–401; March 1953.) "An exact solution is presented for the sound field between two infinite walls when the source is a piston in one of the walls. A method is also outlined for the case of a source having any amplitude distribution which is a function only of radial distance from the center of the source."

534.231 **1877**

The Acoustic Radiation of the Rectangular Piston Membrane—H. Stenzel. (*Acustica*, vol. 2, pp. 263–281; 1952. In German.) General formulas are obtained for calculating the sound field immediately in front of the membrane. The adjacent field is calculated by graphical integration. Field contours for square membranes with different dimension/wavelength ratios are shown. General formulas for the radiation impedance are derived; its two components are calculated and represented graphically for membranes with sides in ratio 1, 2, 5 and 10 to 1.

534.231 **1878**

The Physics of Sound Radiation Pressure—F. Borgnis. (*Z. Phys.*, vol. 134, pp. 363–376; Feb. 6, 1953.) The radiation pressure of a free beam of a plane compression wave incident normally on a plane obstacle is investigated by means of the hydrodynamic impulse law; this method gives a clear physical idea of the forces on the obstacle. The formulas are derived in terms of both Euler and Lagrange variables. Making use of Brillouin's tensor potential a boundary condition for the beam is derived which takes account of the energy exchange between the beam and the ambient medium; this boundary condition leads to a simple relation between the radiation pressure and the difference of the energy densities on the two sides of the obstacle. As a particular case, the radiation pressure on the boundary surface between two immiscible fluids is considered.

534.232:534.321.9 **1879**

Ultrasonic Whistles—A. E. Crawford. (*Research, Lond.*, vol. 6, pp. 106–110; March 1953.) Descriptions are given of generators of resonant-cavity and resonant-wedge type using gas or liquid exciting jets; these are useful for low-power applications such as experiments on ultrasonic propagation in gases, and the emulsification of low-viscosity liquids. Power is provided by a compressor or pump.

534.241:551.510.52 **1880**

Study of the Echoes of Acoustic Waves in the Stratified Region of the Troposphere—G. Eckart. (*Acustica*, vol. 2, pp. 256–262; 1952. In French.) Methods of Bremmer and of Schelkunoff are applied to derive wave equations for sound propagation in a stratified atmosphere. The importance of the terms which render the analogy to em wave propagation incomplete (Eckart & Lienard, Jan. 4) is clearly shown.

534.26 **1881**

On Acoustic Diffraction through an Aperture in a Plane Screen—J. W. Miles. (*Acustica*, vol. 2, pp. 287–291; 1952. In English.) Generalized Fourier transforms are applied to derive equations for the scattered wave. The aperture impedance is expressed in variational form and compared with the Kirchhoff approximation. Reflection from a disk and diffraction at a circular aperture for the case of oblique incidence are considered. (See also 2686 of 1952.)

534.3 **1882**

Experimental Study of the Acoustic, Physiological and Psychophysiological Conditions relating to the Aesthetics of Music—R. Husson. (*Ann. Télécommun.*, vol. 8, pp. 51–72; Feb. 1953.)

534.32:621.396.619.11 **1883**

Perception of Amplitude Modulations—H. Ebel. (*Akust. Beihefte*, no. 4, pp. 246–250; 1952.) Modulation at high frequencies is perceptible at modulation levels lower than that required for perception of low-frequency modulation. Amplitude modulations are less discernible in speech and music than in pure tones. Experimental results agree with those of Zwicker (303 of February) and give an indication of the limiting distortion factor of a loudspeaker for agreeable reproduction.

534.321.9 **1884**

Ultrasonics in Solids—G. Bradfield. (*Research (London)*, vol. 6, pp. 68–79; Feb. 1953.) Thermal vibrations in solids are discussed in relation to physical constants, and the application of ultrasonic and vhf techniques in elasticity measurements is described. Different forms of $BaTiO_3$ transducer are shown and applications of ultrasonics in different technical fields are reviewed.

534.321.9:621.391.6 **1885**

Use of Ultrasonics in Telephony—L. Pimonow. (*Ann. Télécommun.*, vol. 8, pp. 28–30; Jan. 1953.) Measurements have been made of the attenuation of ultrasonic waves propagated in various media, e.g. metal wires, water-filled pipes and air. As a result, equipment has been developed for short-range telephony by means of modulated ultrasonic waves. For two-way operation a suppressed-carrier system may be used; for one-way operation the carrier should be retained. A system is described suitable for conference-room use, with air as the medium of propagation. The carrier-frequency range is 16–100 kc and the radiated power 2–3 W. The receiver is carried on a headband.

534.44 **1886**

Frequency Analysis of Acoustic Phenomena—R. Bierl. (*Akust. Beihefte*, no. 4, pp. 225–235; 1952.) The application of a spectral-function diagram derived from the Fourier integral in filter-circuit and integration methods of waveform analysis is illustrated. Waveforms considered are rectangular and exponential pulses, and sine waves with rectangular and exponential envelopes. The time variation of amplitude and the relative phase of individual frequency components are determined for a limited number of components from the width and position of maxima and the angular velocity of the radius vector of the spectral-function diagram.

534.6:534.321.2 **1887**

Measurement of Harmonic Intervals of the Musical Scale based on the Subjective Sensation of Consonance—M. Barkechli. (*Acustica*, vol. 2, pp. 242–250; 1952. In French.)

534.6:621.395.623 **1888**

The Problem of the Artificial Ear for Cali-

brating Telephone Receivers—I. Barducci. (*Arch. elekt. Übertragung*, vol. 7, pp. 155–157; March 1953.) (See Schiaffino, 1814 of 1952.)

534.62 1889

New Anechoic Chamber of the Technische Hochschule, Karlsruhe—H. Ebel and P. Maurer. (*Akust. Beihefte*, no. 4, pp. 253–256; 1952.) The design and construction of an anechoic chamber are described. By combining wedge-shaped absorbers and damped Helmholtz resonators an absorption factor of 99% for frequencies down to 160 cps is attained. Total thickness of the wall lagging is 61 cm. Absorption characteristics of the material and of the complete chamber are compared.

534.75 1890

Contribution to a Scientific Theory of Single-Channel Transmission of Sound (Part 2)—P. Burkowitz. (*Funk u. Ton*, vol. 7, pp. 10–26; Jan. 1953.) The significance of the amplitude and path-time ratios of reflected and direct sound, and their relation to tonal quality, are discussed (Part 1: 1229 of May).

534.771 1891

The Békésy Audiometer of the Technische Hochschule, Stuttgart—W. Kaiser. (*Akust. Beihefte*, no. 4, pp. 235–238; 1952.) Modifications to the audiometer (Békésy, 3325 of 1949), have been made so that threshold intensities throughout the region of audibility may be accurately determined.

534.771:534.61 1892

The Trend of the Modulation Threshold in the Audible Range—E. Zwicker and W. Kaiser. (*Akust. Beihefte*, no. 4, pp. 239–246; 1952.) Audiometer methods were used to determine the threshold of perceptibility of loudness and pitch variations of pure tones modulated by a 4-cps tone at different modulation levels. Results for cases of defective and also of normal hearing are discussed.

534.84 1893

Diffusion in Architectural Acoustics—W. Furrer and A. Lauber. (*Acustica*, vol. 2, pp. 251–256; 1952. In German.) Analysis of results of measurements shows that the irregularities in the frequency characteristic of a room are not a direct measure of the homogeneity of the sound field, but the mean height of the peaks is a parameter related directly to the "diffusion," or field homogeneity. This parameter has been determined at frequencies around 375 cps and 1650 cps for 11 studios and halls varying in size from 37 m^3 to 20,000 m^3. Two examples are discussed.

534.843 1894

Boundary Conditions for the Acoustic Wave Equation—T. Vogel. (*Acustica*, vol. 2, pp. 281–286; 1952. In French.) The concept of specific normal impedance is discussed in relation to the boundary conditions obtaining for sound propagation in a closed space. Experimental results for a simple case are at variance with this concept. A modified hypothesis is suggested involving two specific impedances, one normal and one tangential to the surface considered.

534.843:534.24.001.57 1895

Reflection of Sound at Surfaces with Periodic Structure—E. Meyer and L. Bohn. (*Akust. Beihefte*, no. 4, pp. 195–207; 1952.) The efficacy of uniformly spaced diffusing elements fixed to plane walls has been investigated on small-scale models, using frequencies in the range 15–60 kc. A diffusion index is calculated, which is the ratio of the reflected sound energy outside the 20°-region of the geometrical reflection to the total reflected energy. The dependence of this index on wavelength, width and depth of the projections and angle of incidence of the sound beam, is discussed for projecting elements of semicircular, triangular and rectangular cross-section in front of (*a*) perfectly absorbing, (*b*) perfectly reflecting surfaces.

534.846.6 1896

Measurement of the Reflecting Power of Ceilings, using Ultrasonics—F. Canac. (*Comp. Rend. Acad. Sci.* (Paris), vol. 236, pp. 467–469; Feb. 2, 1953.) The effectiveness of ceiling treatments such as hollow panels for controlling the reverberation time of halls is investigated by using small-scale models together with proportionately increased frequencies.

621.395.61 1897

Sensitivity of Microphones to Stray Magnetic Fields—L. J. Anderson. (*Trans. Inst. Radio Engrs.*, vol. AU-1, pp. 1–6; Jan./Feb. 1953.) Equipment is described and simple theory is given for (*a*) comparison of various types of microphone with respect to their sensitivity to hum fields, (*b*) measurement of hum fields existing in the neighborhood of microphones. Charts show the hum spectra observed in some particular cases.

621.395.61:534.612 1898

An Experimental Probe Microphone for the Measurement of Sound Pressures—R. B. Archbold. (*P. O. Elect. Eng. Jour.*, vol. 45, part 4, pp. 145–148; Jan. 1953.) Account of the development of a microphone designed primarily for acceptance tests on telephone apparatus. A long narrow tube is coupled to the diaphragm, and the sensitivity/frequency characteristic is smoothed by introducing a small amount of lamb's wool into each end of the tube.

621.395.623.8:621.396.645.371 1899

Portable P.A. [public-address equipment]—E. Griffiths (*Wireless World*, vol. 59, pp. 201–204; May 1953.) Description of equipment giving an output of 8W with <1% distortion for use in small halls. It includes record player, amplifier, loudspeaker and power pack, the whole contained in two cases of about equal size and weight. The negative-feedback amplifier has arrangements for either single-ended or push-pull input, the latter being more suitable for use with the crystal pickup.

621.395.625(083.74):621.396.822 1900

Standards on Sound Recording and Reproducing: Methods of Measurement of Noise, 1953—(Proc. I.R.E., vol. 41, pp. 508–512; May 1953.) Standard 53 IRE 1951.

621.395.625.3 1901

Print-Through Effect in Magnetic Tape for Sound Recording—H. J. Tafel. (*Fernmeldetech. Z.*, vol. 6, pp. 17–24; Jan. 1953.) An experimental investigation of printing through, in adjacent windings of rolled tape, in respect of the intensity and permanence of this magnetization. Its dependence on the temperature, and on the duration, frequency and intensity of the original magnetization is assessed.

621.395.625.3:534.76 1902

A Practical Binaural Recording System—O. C. Bixler. (*Trans. Inst. Radio Engrs.*, vol. AU-1, pp. 14–22; Jan./Feb. 1953.) Modifications to commercial tape recording and reproducing equipment are described resulting in a simple stereophonic system comprising two microphones and two loudspeakers, each channel using half the tape. Various applications are indicated, including court-room reporting. Broadcasting tests were made by connecting one microphone to an A.M. transmitter and the other to a F.M. transmitter, the listener using A.M. and F.M. receivers simultaneously.

621.395.92:621.314.7 1903

Transistorized Hearing Aids—J. D. Fahnestock. (*Electronics*, vol. 26, pp. 154–155; April 1953.) Circuit and performance details are given of various commercially available hearing aids using (*a*) subminiature valves and transistors or (*b*) transistors exclusively. Only junction-type transistors are used, with grounded-emitter circuits. Noise level is discussed. Some models have facilities for switching from microphone to a telephone pickup coil.

681.84 1904

Piezo-electric Crystal Pick-ups—S. Kelly. (*Jour. Brit. IRE*, vol. 13, pp. 161–170; March 1953.) Bimorph crystal elements are used, usually torsional in the case of Rochelle salt, and always benders in the case of $BaTiO_3$. Design methods make use of equivalent electrical circuits. Needle-tip impedance can be reduced to obtain satisfactory tracking at a needle pressure of <5 gm. Frequency response can be levelled from 20 cps to 20 kc, using simple equalizers. Wear and distortion are discussed. Details are given of a two-stylus pickup for 78- and $33^1/_3$-rpm records.

AERIALS AND TRANSMISSION LINES

621.315.2.015.7 1905

Theory of Pulse Technique for Coaxial Cables—P. Behrend. (*Z. angew. Phys.*, vol. 5, pp. 61–64; Feb. 1953.) Analysis is given for the deformation experienced by rectangular or half-sine-curve pulses traversing a coaxial cable, the transfer function of the cable being assumed to be a subfunction of a Laplace transform. Results are shown in graphs.

621.392.21.09 1906

Physical Explanation of the Surface Wave on a Dielectric-Coated Line—P. J. M. Clavier. (*Câbles & Trans.* (*Paris*), vol. 7, pp. 34–38; Jan. 1953.) The propagation of a surface wave is treated as a series of reflections alternately at the metal surface and the dielectric/air interface. By analogy with waveguide propagation an E_0 mode is defined. Below the limiting frequency for progagation along the line, energy is radiated and the mode analogy is destroyed. The relation between angle of incidence on the reflecting surfaces and the phase and group velocities is illustrated and the variation of the latter as a function of frequency is shown graphically.

621.392.26:621.315.61 1907

Propagation in Waveguides filled Longitudinally with Two or More Dielectrics—L. G. Chambers. (*Brit. Jour. Appl. Phys.*, vol. 4, pp. 39–45; Feb. 1953.) Discussion of the propagation of pure TE and TM modes in inhomogeneously filled rectangular and circular waveguides, summarizing the results of research work published since 1943. 16 references.

621.396.67 1908

An Alternative Method of solving Hallén's Integral Equation and its Application to Antennas near Resonance—R. King. (*Jour. Appl. Phys.*, vol. 24, pp. 140–147; Feb. 1953.) Hallen's equation is separated into two equations for the in-phase and quadrature-phase current components respectively; these are solved by an iteration method. For values of aerial electrical half-length near odd multiples of $\lambda/4$, at least a third-order solution is required for the accurate determination of in-phase current and conductance. Conductance values for a range of radii are evaluated using the third-order formula; comparison with experimental results indicates that for antennas near resonance, just as for very short and very long antennas, account must be taken in the iteration procedure of both the current components.

621.396.67 1909

Power Gain of Curtain Arrays of Aerials—P. Hammond. (*Wireless Eng.*, vol. 30, pp. 108–111; May 1953.) The directional concentration of power by systems comprising uniformly spaced antennas carrying equi-phase currents is investigated using the concept of radiation resistance. The field due to an infinitely long filament is first considered and a solution is obtained in terms of Bessel functions of zero order and of the first and second kinds. From this the power output can be determined for a single antenna and for combinations of various numbers of antennas in arrays. For a spacing of about $3\lambda/4$ between elements, the power gain passes through a maximum, its value being then about 50% greater than the value for $\lambda/2$ spacing. This maximum is also obtained for finite-height antennas and for arrays having depth as well as width.

621.396.67:621.315.612.4:621.396.611.4 1910
Quasi-Degenerated Modes in High-ϵ Dielectric Cavities—Schlicke. (See 1942.)

621.396.67:621.396.9 1911
Moulded Plastic Radar Scanners and Stressed Components—(*Engineering* (*London*), vol. 175, pp. 28–29; Jan. 2, 1953.) The 14-ft. radar reflector described consists of a sandwich-like construction having outer layers of asbestos-fibre material impregnated with phenolic resin and a resin-impregnated paper-honeycomb core. It is as strong as, stiffer than, and weighs half as much as a reflector made of Al alloy; resistance to weathering is good.

621.396.677 1912
Reception Diagrams of Rhombic Aerials in a Vertical Plane—J. Dufour. (*Tech. Mitt. Schweiz. Telegr.-TelephVerw.*, vol. 31, pp. 65–72; March 1, 1953. In French.) Measurements were made of the received signal strength on the ground, at Châtonnaye, using an airborne transmitter. Results obtained with two slightly different rhombic aerials were compared with the theoretical values for frequencies of 5, 10, 15, 20, 25 and 30 mc. Good agreement was found. There is a strong ground reflection at 6–7 km from the station, giving rise to marked interference for angles of incidence $<8°$ from the horizontal.

621.396.677 1913
A Theory of Plane Reflectors in Microwave Antenna Systems—H. L. Knudsen and M. Andreasen. (*Trans. Dan. Acad. Tech. Sci.*, no. 3, pp. 3–57; 1952.) Values are calculated for the field reflected from an apparently circular reflector illuminated by a point radiator; they differ considerably from results obtained by Aasma. For the case of a finite-size radiator, the calculations are based on the assumption of rectangular radiator and reflector apertures. Curves are presented from which the reflected field and the gain can be read directly or determined by a combination of two readings. Errors introduced by the simplifying assumptions are discussed. Fresnel integrals and related functions are used in the analysis; relevant tables and curves are given.

621.396.677:621.392.26 1914
Second-Order Beams of Slotted-Waveguide Arrays—H. Gruenberg. (*Canad. Jour. Phys.*, vol. 31, pp. 55–69; Jan. 1953.) Second-order beams due to slot offset have escaped detection because they are not located in the plane containing the waveguide axis, where the radiation pattern is usually measured. The patterns of the second-order beams for a waveguide with staggered slots are analyzed by considering an approximately equivalent arrangement of two parallel linear arrays separated by twice the weighted mean offset. The second-order beams can be suppressed by using (*a*) an array with all slots aligned and appropriate corrugations in the waveguide wall, or (*b*) a conventional waveguide array fitted with a horn incorporating a parallel-plate section (3343 of 1952).

621.396.677:621.392.26 1915
Theory of Waveguide-Fed Slots radiating into Parallel-Plate Regions—L. Lewin. (*Jour. Appl. Phys.*, vol. 24, p. 232; Feb. 1953.) Simpler alternative expressions are derived for integrals introduced in the original paper by Gruenberg (3343 of 1952).

621.396.677.029.62 1916
Theory of Artificial Slot Antennas—A. Ataman. (*Bull. Tech. Univ. Istanbul*, vol. 4, pp. 71–89; 1951. In English.) The term "artificial slot" is applied to a radiator comprising a dielectric rod of rectangular cross-section, lying on a conducting surface and having two sides covered with metal plates between whose upper edges an electric field is established. Such devices are useful for the frequency range 100–300 mc, and may have advantages over a true slotted aerial for use in high-speed aircraft. Analysis is given, treating the arrangement as a uniformly loaded transmission line. A determination is made of the properties of the dielectric required to give satisfactory radiation efficiency; high permeability and low losses are desirable. A numerical calculation is made for a $\lambda/2$ aerial to operate at 200 mc with a slot width of 2 cm; for a material with a dielectric constant of 10 the slot depth is 0.5 cm.

621.396.677.5:621.318.132 1917
Receiving Properties of a Wire Loop with a Spheroidal Core—J. R. Wait. (*Canad. Jour. Tech.*, vol. 31, pp. 9–14; Jan. 1953.) The relative gain of a loop antenna with spheroidal core is calculated, assuming core losses to be negligible, an assumption which is justified for ferromagnetic ceramics at frequencies of the order of 200 kc or lower. The magnetic properties of modern ferromagnetic materials are used to good advantage when the spheroid is highly elongated or very much flattened.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.22/.23:621.315.512 1918
Internal Alternating-Current Resistance and Reactance of Conductors of Circular Cross-Section—P. M. Prache. (*Câbles & Trans.* (*Paris*), vol. 7, pp. 28–33; Jan. 1953.) A numerical table is presented for simple and accurate calculations for solid conductors and for thin or thick tubes. Proximity effect is neglected.

621.3.014/.016:621.3.012.1 1919
Vector Representation in the Time Plane of Electrical Potential Difference, Current and Power varying Sinusoidally in Time—G. Kouskoff. (*Rev. gén. Elect.*, vol. 62, pp. 86–90; Feb. 1953.) The varying magnitudes are considered, according to a method originally used by Fresnel, as the resultant of two imaginary vectors of equal and constant length, rotating at the same rate in opposite senses. The imaginary plane containing these vectors is called the "time plane." The impedances of a circuit to the two vector components of a sinusoidally varying current are defined, and the method of representing power, i.e. the product of voltage and current, is explained.

621.3.015.7:621.387.4 1920
A New Pulse-Analyzer Design—C. W. Johnstone. (*Nucleonics*, vol. 11, pp. 36–41; Jan. 1953.) Description, with detailed circuit diagrams, of a single-channel analyzer which serves as the basic design for multichannel analyzers, details being given of a 10-channel analyzer. Positive pulses are accepted in the amplitude range 3–103 V plus the sum of channel widths. The width and position of the channels can be changed quickly without requiring recalibration. The design is simple and relatively few tubes are required.

621.316.86:537.312.6 1921
Resistor Temperature Coefficients—F. A. Paul. (*Tele-Tech*, vol. 12, pp. 52–53, 120; Jan. 1953.) Tests were carried out on four types of pyrolytic-carbon-film resistor and one type of metal-film resistor, several samples of each type, of different resistance ratings, being used. The results obtained for the temperature range from -30) to $+120°$ C are shown graphically. The temperature coefficients of the carbon-film resistors are considerably lower than those of carbon-composition resistors. The coefficients are negative and much greater than those of wire-wound resistors, though this is not the case for borrocarbon-film resistors [2365 of 1951 (Grisdale *et al.*)].

621.318.4.011 1922
Determination of Voltage, Current, and Magnetic Field Distributions together with the Self-Capacitance, Inductance and H.F. Resistance of Single-Layer Coils—A. E. S. Mostafa and M. K. Gohar. (Proc. IRE, vol. 41, pp. 537–547; May 1953.) A theoretical investigation is made which is applicable to coils of intermediate lengths. The number and magnitudes of the harmonics are calculated, and their effect on the self-capacitance is indicated. Experimental results obtained by other workers support the theoretical results.

621.318.42:538.221 1923
Ferrite Core Inductors—H. A. Stone, Jr. (*Bell Sys. Tech. Jour.*, vol. 32, pp. 265–291; March 1953.) Design of inductors for communication equipment is considered particularly from the point of view of the best use of the properties of ferrites. Advantages of ferrite compared with metal cores for high-frequency low-power applications include (*a*) higher Q values, (*b*) smaller volume, and (*c*) greater ease of adjustment.

621.318.42:621.311.6 1924
Miniaturization of Airborne Filter Chokes—W. E. Tanner. (*Electronics*, vol. 26, pp. 180–183; April 1953.) A calculation chart is given for chokes using 7-mil laminations of 4% Si steel with square stack; use of the chart is illustrated by numerical examples.

621.319.4 1925
The Power Factor and Capacitance of Mica Capacitors at Low Frequencies—P. R. Bray. (*Jour. Sci. Instr.*, vol. 30, pp. 49–51; Feb. 1953.) Experimental results confirm an empirical law of the form af^m (m negative and fractional) for the power factor over the range 1–30 cps at normal temperatures. They also confirm a relation between frequency, power factor and capacitance. Power factors as high as 0.01 have been found at 1 cps, corresponding to increases of nearly 2% in capacitance over the values at higher frequencies.

621.392.2:621.3.015.3 1926
Calculation of Transient Phenomena in Electrically Short Networks by means of Fourier Integrals—E. Adams. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, vol. 7, pp. 1–10 and 36–40; Jan./April 1953.) Fourier integrals are applied to the determination of the frequency spectra of (*a*) periodic functions, (*b*) nonperiodic functions including step functions. The results are used in discussion of the distortion and initial conditions in short networks to which an impulse voltage is applied.

621.392.5 1927
Design of Practical Linear Networks with Optimum Signal/Noise Ratios—K. Halbach. (*Helv. Phys. Acta.*, vol. 26, pp. 65–74; Feb. 15, 1953. In German.) A general method is described for the design of networks for which the relation between the input and output signals is represented by linear equations, the frequency spectra of signal and noise being assumed known. The application of the method of analysis in the design of the optimum ionization-chamber amplifier is illustrated, and the signal/noise ratios attainable are compared with those for an *RC* discriminating network and a delay-line clipping circuit.

621.392.5 1928
Applications of *T* Networks—M. Alixant. (*Radio Tech. Dig., Édn franc.*, vol. 6, pp. 301–316; 1952.) Adaptation of paper by Lavagnino (3032 of 1952) incorporating additional material. 53 references.

621.392.5 1929
Synthesis of Transfer Functions with Poles Restricted to the Negative Real Axis—L. Weinberg. (*Jour. Appl. Phys.*, vol. 24, pp. 207–216; Feb. 1953.) A procedure for the synthesis of general *RC* transfer functions by means of unbalanced networks is described. The transfer function need not be minimum-phase, but may have zeros anywhere in the complex plane except on the positive real axis. Use is made of techniques described by Guillemin (2462 of 1949), but the problem is simplified by use also

of a network theorem due to Adler, enabling zero shifting to be performed in two directions from within the total network. In an example worked out, 26 network elements are used, whereas the Guillemin procedure would require 66.

621.392.5:538.652:546.289 1930

Hall-Effect Modulators and "Gyrators" employing Magnetic-Field-Independent Orientations in Germanium—W. P. Mason, W. H. Hewitt and R. F. Wick. (*Jour. Appl. Phys.*, vol. 24, pp. 166–175; Feb. 1953.) The application of the Hall effect in Ge to the measurement of magnetic flux has been described previously [Pearson, 3172 of 1948]. Further applications are now described to (*a*) a product modulator, in which the current corresponding to one signal is passed through the Ge crystal while that corresponding to the other is used to produce the magnetic field, and (*b*) a circuit element with nonreciprocal transmission properties, or gyrator. In these applications a crystal orientation is used for which the transverse magnetoresistance effects are zero and the variation of the Hall-effect coefficient with flux up to 20,000 gauss is $\not> 2\%$. The experimental determination of this orientation is described.

621.392.52 1931

The Development of the Electrical Filter by K. W. Wagner's Methods—T. Laurent. (*Arch. elekt. Übertragung*, vol. 7, pp. 126–135; March 1953.) Wagner's method of filter design, which is based on consideration of the filter as a ladder of matched reflection-free sections, is compared with methods developed by Zobel, Cauer and others; it is concluded that the principles laid down by Wagner have lost none of their original validity or usefulness.

621.392.52 1932

Some Designs of Ladder Filter. Results Obtained and Methods of Calculation—D. Starynkevitch. (*Câbles & Trans.* (*Paris*), vol. 7, pp. 3–15; Jan. 1953.) The principle of matching half-sections and transforming impedances in designing filters with a minimum number of components is illustrated in two examples: a narrow-band band-pass filter for 612 kc, and a filter with a 4-mc pass band centered on 10 mc. Formulas relating to the transformation of T-type quadripoles to Π-type and vice versa, and to various types of quadripoles for impedance matching, are collected in appendices.

621.392.52:[621.392.26<621.315.212 1933

U.H.F. Band-Pass Filters—J. C. Simon and G. Broussaud. (*Ann. Radioélect.*, vol. 8, pp. 3–19; Jan. 1953.) A direct method is developed for designing waveguide and coaxial-line filters; the characteristics of a 3-cavity filter are calculated. Transverse obstacles comprising sets of thin pins are shown to be better than diaphragms, windows or single pins from the point of view of losses and of manufacture and adjustment. Cavities with Q values of 100–200 can easily be obtained using four pins of diameter 1–2 mm. Design charts are given, and several types of filter are described and illustrated.

621.392.52.011.21 1934

Admittance and Transfer Function of a Multimesh Resistance-Capacitance Filter Network—B. K. Bhattacharyya. (*Indian Jour. Phys.*, vol. 26, pp. 563–574; Nov. 1952.) The impedance function of an n-mesh RC network is expressed as a recurring continued fraction which is then resolved into the quotient of two polynomials, the admittance function being the reciprocal of this quotient. The transfer function is then determined. The corresponding formulae for an n-mesh CR network are derived.

621.392.52.029.64:621.396.611.4 1935

Determination of the Bandwidth of Band-Pass Filters for Centimetre Waves—E. Ledinegg and P. Urban. (*Arch. elekt. Übertragung*, vol. 7, pp. 99–105; Feb. 1953.) Theory developed previously (1115 of 1951) is used to analyze the band-pass-filter operation of an arbitrary arrangement of coupled cavity resonators. A detailed treatment is given for the system comprising two stagger-tuned coupled resonators.

621.392.53 1936

Spectrum Equalization—G. G. Gouriet. (*Wireless Eng.*, vol. 30, pp. 112–123; May 1953.) The transfer characteristics of nearly all forms of linear quadripoles can be equalized as regards both amplitude and phase by adding to the response function its successive time derivatives or integrals. The method may be applied to systems for which the equivalent passive quadripole is not realizable, e.g. a scanning aperture. Application of the method to servo-control and television is discussed.

621.392.54 1937

U.H.F. Magnetic Attenuator—F. Reggia. (*Radio & Telev. News, Radio-Electronic Eng. Sec.*, vol. 49, pp. 12–14, 24; April 1953.) Construction details and performance curves are given for attenuators of the type previously described (1237 of 1952).

621.392.54:621.392.26 1938

Cut-Off Attenuators for Waveguide Circuits—G. Klätte. (*Fernmeldetedh. Z.*, vol. 6, pp. 86–89; Feb. 1953.) An equivalent-circuit analysis and report of measurements on short and long cut-off sections providing attenuation respectively less than and greater than 15 db. Beyond a certain length the reduction in attenuation due to reflections at the junction is constant, so that calibration measurements on long sections are unnecessary.

621.396.6+621.385]:629.13.05 1939

Reliability of Airborne Electronic Components—B. G. Bromber and R. D. Hill, Jr. (PROC. IRE, vol. 41, pp. 513–516; May 1953.) An analysis is made of typical failures. Reliability may be improved by modifications to present-day components and by substitution of less vulnerable elements, such as transistors and magnetic amplifiers.

621.396.611.1 1940

Forced Oscillations with Nonlinear Restoring Force—C. Hayashi. (*Jour. Appl. Phys.*, vol. 24, pp. 198–207; Feb. 1953.) Both transient and steady-state oscillations are studied. The original differential equation characterized by a nonlinear term is transformed to the following first-order differential equation:—

$$dy/dx \quad Y(x, y)/X(x, y).$$

Following Poincaré and Bendixson, the singularities and the integral curves of this equation are discussed, the former being correlated with the steady states and the latter with the transient states of oscillation. The integral curves yield the relation between the initial conditions and the steady-state solutions. Fundamental and subharmonic oscillations of order $\frac{1}{3}$ are investigated. The theoretical results are compared with measurements made on a circuit containing a saturable iron core; satisfactory agreement is found.

621.396.611.1 1941

Response of a Series of Resonant Circuits to an Applied E.M.F. whose Frequency is a Linear Function of Time—J. Marique. (*Ann. Télécommun.*, vol. 8, pp. 43–50; Feb. 1953.) A formula is derived for the response of two or three RLC circuits in series to an e.m.f. of unit amplitude represented by the expression $e(t) = \exp.\, j(\omega_0 t + \epsilon t^2)]$, where ϵ is half the frequency-sweep speed. The fluctuations in the response curve of a single circuit are very much reduced when several circuits are connected in series. The increase of selectivity obtainable is investigated, and frequency-sweep speeds permissible for one, two and three circuits are determined.

621.396.611.4:621.315.612.4:621.396.67 1942

Quasi-Degenerated Modes in High-ϵ Dielectric Cavities—H. M. Schlicke. (*Jour. Appl. Phys.*, vol. 24, pp. 187–191; Feb. 1953.) A theoretical and practical investigation is made of resonators comprising solid circular cylinders composed of materials such as titanates, having permittivities of 2500–10,000, with or without metallization of the surfaces. As a consequence of the high permittivities, cavity-resomator techniques can be used in the vhf range. For cylinders with the ends not metallized, quasi-degenerate TE modes can be realized. Tuning can be effected by inserting magnetic rods or by applying conductive disks externally. A design for an antenna developed from the $TE_{010}^{\infty\infty}$ mode resonator comprises a dielectric spiral coiled round a magnetic rod. Practical disadvantages inherent in these high-permittivity dielectrics are indicated.

621.396.615:621.316.729 1943

The Method of Analytical Signals. Study of the Synchronization of Amplitude-Limited Oscillators—J. Daguet. (*Ann. Télécommun.*, vol. 8, pp. 2–10; Jan. 1953.) A method based on that of Ville (1020 of 1950) is used to determine the frequency limits within which synchronization can be effected for a single-tuned self-limiting tube oscillator. The synchronization range is given by $2\Delta\omega/\omega \leq E_0/QE$, where $\Delta\omega$ is the difference between the frequency of the synchronization voltage E_0 and the natural frequency ω of the oscillator, and E is the amplitude of the oscillation in the steady state. The phase angle between the two voltages varies from 0 for $\Delta\omega = 0$ to 90° for $\Delta\omega = \Delta\omega_{\text{max}}$. A system with wide synchronization range is discussed; this incorporates a phase discriminator whose output, proportional to the phase angle between synchronization and oscillator voltages, is applied to a reactance tube which in turn controls oscillator frequency.

621.396.615:621.392.4/.5 1944

The Use of Admittance Diagrams in Oscillator Analysis—H. J. Reich. (PROC. IRE, vol. 41, pp. 522–528; May 1953.) A digest of the available information on this technique.

621.396.615:621.396.822:529.786 1945

Effect of Background Noise on the Frequency of Tube Oscillators. Ultimate Accuracy of Electronic Clocks—A. Blaquière. (*Ann. Radioélect.*, vol. 8, pp. 36–80; Jan. 1953.) Fuller account of work noted previously (2163 of 1952, 968 of April, and back references).

621.396.615.17 1946

How to design Bistable Multivibrators—R. Pressman. (*Electronics*, vol. 26, pp. 164–168; April 1953.) The design procedure described is intended to ensure reliable operation in spite of deviations of resistors from nominal values, variations of supply voltage, and deterioration of tube-cathode emission. Formulas are given for determining the appropriate values of coupling capacitance. Triggering networks for dealing with pulses of different amplitude and duration are shown.

621.396.645 1947

The Effective Bandwidth of Video Amplifiers—F. Tischer. (*Kungl. tek. Högsk. Handl., Stockholm*, no. 63, pp. 1–32; 1953.) Expanded version of 3050 of 1952.

621.396.645:621.396.615.142.2 1948

Fluctuation Noise in a Microwave Superregenerative Amplifier—T. S. George and H. Urkowitz. (PROC. I.R.E., vol. 41, pp. 516–521; May 1953.) The theory of operation of a superregenerative amplifier is reviewed; Rice's representation of noise current by a Fourier series (2169 of 1945 and back references) is used in the expression for the amplifier output. A study is made of the particular conditions when a reflex klystron is used in the amplifier; expressions are derived for the noise figure in terms of the klystron parameters. The overall noise figure can be reduced by introducing beam current, with its associated shot noise, during the quiescent interval.

621.396.645.018.424 1949

Calculation of Compensated Wide-Band Amplifiers—H. Laett. (*Tech. Mitt. schweiz. Telegr.-Teleph. Verw.*, vol. 31, pp. 72–80; March 1, 1953. In German and French.) Loss of amplification with increasing frequency, due to the output capacitance of the amplifier tube, is compensated by including inductance in the circuit. Three arrangements are discussed, namely parallel compensation, series compensation and double compensation. Abacs to facilitate calculations are presented.

621.396.645.35 1950

D.C. Amplifiers with Negative Feedback—A. Ehmert. (*Z. angew. Phys.*, vol. 5; pp. 24–33; Jan. 1953.) Methods are described for automatically compensating the input impedance of current amplifiers or the input admittance of voltage amplifiers. Details are given of the design and performance of several circuits for measurement purposes, in which very high input impedance with low capacitance is achieved.

621.396.654.371:621.317.733 1951

The Wien Bridge as a Frequency-Determining Element in Selective Amplifiers—H. J. Stöhr. (*Funk u. Ton*, vol. 7, pp. 27–33; Jan. 1953.) Analysis of the balance condition in a bridge circuit suppressing the amplifier feedback voltage at a fixed frequency.

621.396.662.029.63 1952

Coaxial–Cavity Tuning Element for U.H.F.—(*Radio & Telev. News, Radio-Electronic Eng. Sec.*, vol. 49, pp. 15–16, 30; April 1953.) The design is described of a unit which can be used in television tuners and test instruments for the 470–890 mc band. The unit consists essentially of a metal cylinder with an axial conductor having a gap at the center to form a variable capacitor for tuning. An accurately machined low-loss dielectric tube separates core and cylinder and serves as a close-fitting guide for the tuning plunger. Applications of the unit in a signal generator, a frequency-sweep generator, a wavemeter, and a grid-dip meter are described briefly.

621.396.665:621.316.86 1953

A.G.C. by Means of Miniature NTC Resistors with Heating Element—G. H. Schouten. (*Electronic Applic. Bull.*, vol. 12, pp. 33–37; Feb. 1951.) The negative-temperature-coefficient resistors are surrounded by the heating element and enclosed in an evacuated glass tube, the resistor being included in the input circuit of an amplifier and the heating element in the input circuit to obtain efficient a.g.c. Practical circuit details are given and operating characteristics of the NTC resistors are shown graphically.

621.397.645.029.62 1954

Fundamental Problems of H.F. and I.F. Amplifiers for TV Reception: Part 3—Feedback and Practical Considerations following on the Theory—A. G. W. Uitjens. (*Electronic Applic. Bull.*, vol. 12, pp. 41–64; Feb. 1951.) A detailed discussion is presented of the effects in television amplifiers of feedback (*a*) via the cathode-grid capacitance, (*b*) via the anode-grid capacitance, (*c*) via the anode-cathode capacitance. The necessary formulas are derived for an amplifying stage in which all three types of feedback are present, so that their interactions can be investigated theoretically. The detailed discussion is, however, confined to systems in which only one form of feedback is present. The theory is illustrated by practical examples in amplifier design.

621.392.52 1955

Filter Design Data for Communication Engineers—J. H. Mole. (Book Review.) Publishers: E. & F. N. Spon, London, 252 pp., 63s. (*Wireless Eng.*, vol. 30, pp. 129–130; May 1953.) The book is intended to supplement existing textbooks by providing charts, tables and formulas in convenient form; it is concerned mainly with image-parameter (Zobel) filters.

GENERAL PHYSICS

530.12 1956

The Relativistic Electromagnetic Equations in a Material Medium—N. W. Taylor. (*Aust. Jour. Phys.*, vol. 6, pp. 1–9; March 1953.) The general relativistic e.m. equations for a material medium are expressed in the form of a single four-vector density equation. The field tensor has six different complex components instead of three, as in the case of a free medium. The classical equations are obtained by separating the real and imaginary parts.

530.145:621.314.7 1957

Transistors: Theory and Application: Part 2—Energy Levels in Transistor Electronics—A. Coblenz and H. L. Owens. (*Electronics*, vol. 26, pp. 138–141; April 1953.) A simplified exposition of some of the fundamentals of quantum mechanics relevant to an understanding of transistor operation.

530.145.6:537.311.33:621.396.11 1958

Some Particular Features of Tropospheric Propagation and their Analogies in Wave Mechanics—Ortusi. (See 2112.)

537.122:621.317.318 1959

A Precise Determination of the Charge of the Electron from Shot Noise—L. Stigmark. (*Ark. Fys.*, vol. 5, pp. 399–426; Oct. 26, 1952.) The electron charge e is calculated from the shot fluctuations of saturation current in a specially constructed diode connected across a tuned circuit resonant at 600 kc. From precision measurements of circuit parameters and the mean-square value of shot noise, the final result obtained is $e=(4.797\pm 0.012)\times 10^{-10}$ esu.

537.311.33 1960

Electron Spin in Semiconductors—E. A. Guggenheim. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 121–122; Jan. 1, 1953.) Theory is developed showing that, starting from equilibrium electron distribution expressed in terms of absolute activity, it is possible correctly to take account of electron spin.

537.523/.525 1961

High–Frequency Discharges in Gases—G. D. Morgan. (*Sci. Progr.*, vol. 41, pp. 22–41; Jan. 1953.) Discussion of the main characteristics and the mechanism of HF discharges, indicating their practical applications. 70 references.

537.525:537.533 1962

Field Emission of Electrons in Discharges—F. L. Jones and E. T. de la Perrelle. (*Proc. Roy. Soc. A*, vol. 216, pp. 267–279; Jan. 22, 1953.) Substantial emission of 10^4 to 10^5 electrons/second was obtained from cathodes of oxidized Ni and W with electric fields of about 100 kV/cm. By relating this emission to the electric field by the Fowler-Nordheim equation, the effective work function and emitting area of the source of the electrons for oxidized Ni and W were estimated as about 0.5 eV and 10^{-13} cm^2 respectively. These results are discussed.

537.533 1963

Field Electron Emission and Work Function of Crystal Surfaces—M. Drechsler and E. W. Müller. (*Z. Phys.*, vol. 134, pp. 208–221; Jan. 20, 1953.) The field close to the surface is determined by measurements on models in an electrolyte tank. Formulas are derived for the density of the field-emission current. These are verified by field-electron-microscope observations on W crystals.

537.562:538.122 1964

Plasma in a Magnetic Field—A. Schlüter. (*Ann. Phys., Lpz.*, vol. 10, pp. 422–429; July 15, 1952.) The results obtained by considering plasma (*a*) from the electron-motion point of view, (*b*) on the basis of kinetic theory, are discussed and reconciled.

537.562:538.56 1965

The Relation between Various Types of Plasma Oscillation—A. Schlüter. (*Ann. Phys. Lpz.*, vol. 10, pp. 418–421; July 15, 1952.) A dispersion relation is derived which applies to plasma oscillations, electron and ion oscillations of the Larmor type and to magnetohydrodynamic waves.

538:621.3.01 1966

Definition of Magnetic Quantities—G. Oberdorfer. (*Arch. elekt. Übertragung*, vol. 7, pp. 136–142; March 1953.) A critical examination is made of the nomenclature hitherto used for magnetic quantities, and proposals are made for definitions which fit in with modern ideas without unnecessarily disturbing established usage. Precise definitions are given of field strength, polarization, magnetization, susceptibility and magnetizability.

538.221 1967

Some Magnetic Properties of Metals: Part 5—Magnetic Behaviour of a Cylindrical System of Electrons for All Magnetic Fields—R. B. Dingle. (*Proc. Roy. Soc. A* (*London*), vol. 216, pp. 118–142; Jan. 7, 1953.) The Schrödinger equation is solved for an electron moving in a uniform magnetic field H, the boundary of the system being a cylinder with its axis along the direction of the field. Two entirely different types of wave function are possible, one type leading to the small Landau diamagnetism of large systems (Part 1: 2489 of 1952), the other to the larger diamagnetism of small systems (Part 4: 1005 of April). Taking account of the occupied states of both types, the steady contributions to the magnetic susceptibility are determined for low, high and intermediate temperatures.

538.221 1968

Collective Electron Ferromagnetism: a Generalization of the Treatment and an Analysis of Experimental Results—K. L. Hunt. (*Proc. Roy. Soc. A* (*London*), vol. 216, pp. 103–117; Jan. 7, 1953.) A term involving the fourth power of the magnetization is introduced into Stoner's expression for the energy associated with magnetization. If parameter values are suitably chosen, the properties of Ni show a high degree of coordination, and the magnetization/temperature curves for Ni-Co and Ni-Cu alloys can be interpreted simply.

538.3 1969

Tentative Theory of Nonlinear Electrodynamics: Part 2—K. Bechert. (*Ann. Phys., Lpz.*, vol. 10, pp. 430–448; July 15, 1952.) Part 1: 618 of 1951.

538.561:537.533/.534 1970

Production of Radio Waves by means of the Tcherenkov Effect—C. A. Klein. (*Ann. Télécommun.*, vol. 8, pp. 38–42; Feb. 1953.) The Tcherenkov effect consists in the emission of em radiation by a particle traveling through a medium with a velocity greater than that of light. The essential results of Frank & Tamm's theory of the effect (*Compt. Rend. Acad. Sci.* (*URSS*), vol. 14, pp. 109–114; 1937) are quoted and references given to accounts of experimental verifications of the theory, using electrons and also protons. The possibility of producing em waves in the range between the far infrared and uhf by means of the effect in a narrow channel is discussed. It is concluded that intense sources of em waves should result from the use of powerful beams consisting of packets of electrons. Simple arrangements should suffice for cm and mm waves, but difficulties will certainly be experienced in the case of still shorter waves.

538.566:535.42 1971

A Rigorous Treatment of the Diffraction of Electromagnetic Waves by a Slit—R. Müller and K. Westpfahl. (*Z. Phys.*, vol. 134, pp. 245–263; Feb. 6, 1953.) Wave-polarization components parallel to and perpendicular to the slit are considered. An integral equation for the electric field strength in the slit is found by the method of Levine & Schwinger (83 and 1897 of 1950), and an exact solution is obtained by developing the kernel in a power series of the product of wave number and slit width.) The far-field intensity and the transmission factor of the slit are calculated and compared with the results of Morse and Rubenstein (905 of 1939). The relation of the work to that of Sommerfeld (Vorlesung über theoretische Physik, Bd IV, Optik, §39) and of Groschwitz and Hönl (2183 of 1952) is discussed.

538.566:535.42 1972

The Three-Dimensional Problem of the Diffraction of Monochromatic Electromagnetic Waves—D. Z. Avazashvili. (*Compt. Rend. Acad. Sci.* (*URSS*), vol. 28, pp. 29–32; Jan. 1, 1952. In Russian.) The distribution of the em field in space for a given distribution of field at a boundary surface is discussed, and conditions are established for the existence of a solution obtained previously (*Trudy Tbilisskovo Matematicheskovo Instituta*, vol. 8, p. 109; 1940).

538.566:621.318.423:621.385.029.6 1973

The Propagation of Electromagnetic Waves along a Conical Helix with Variable Pitch—G. Hellgren. (*Chalmers tek. Högsk Handl.*, no. 130, 13 pp.; 1953.) Theoretical analysis. Expressions are derived for the reflection coefficient, which may be practically zero, and for the impedance characteristics.

538.63 1974

Galvanomagnetic Effects in Conductors—D. K. C. MacConald and K. Sarginson. (*Rep. Progr. Phys.*, vol. 15, pp. 249–274; 1952.) A review of the Hall effect and magnetoresistance in conductors, with primary reference to metals (including ferromagnetic materials); semiconductors are mentioned briefly. A physical survey of the whole field is first made; in the second part a more mathematical treatment is given of some of the major problems involved. Over 60 references.

538.632 1975

Hall Effect—O. Lindberg. (Proc. I.R.E., vol. 41, p. 507; May 1953.) Correction to paper noted in 713 of March.

538.652 1976

Magnetostriction—E. W. Lee. (*Sci. Progr.*, vol. 41, pp. 58–77; Jan. 1953.) Discussion of magnetostriction phenomena in single crystals and in multicrystal materials. Three effects are distinguished: (*a*) form effect, (*b*) "forced" magnetostriction, (*c*) spontaneous linear magnetostriction.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5:621.396.9 1977

On the Nature of the Decay of a Meteor Trail—J. Feinstein. (*Proc. Phys. Soc.* (*London*), vol. 65, p. 741; Sept. 1, 1952.) Critical discussion of the theories of Kaiser and Closs (1108 of 1952) and of Greenhow (2493 of 1952).

523.5:621.396.9 1978

On the Decay of Radio Echoes from Meteor Trails—T. R. Kaiser and J. S. Greenhow. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 150–151; Feb. 1, 1953.) Discussion shows that observed phenomena are explicable by the theories of Kaiser and Closs (2208 of 1952) and of Greenhow (2493 of 1952). Many points of disagreement between observation and Feinstein's theory (1885 of 1951 and 1977 above) are noted.

523.5:621.396.9 1979

An Experimental Study of Radio Reflections from Meteor Trails—R. L. Closs, J. A. Clegg and T. R. Kaiser. (*Phil. Mag.*, vol. 44, pp. 313–324; March 1953.) The dependence of the echoes on the polarization of the incident wave has been investigated and the existence of resonance scattering with transverse polarization verified.

523.72 1980

Thermal Theories of the High-Intensity Components of Solar Radio-Frequency Radiations—J. H. Piddington. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 97–104; Feb. 1, 1953.) Thermal theories of the origin of enhanced RF radiation from regions which occasionally attain brightness temperatures of 10^{10} deg K or more are discussed. None of the theories is found capable of explaining the available observational data. It is concluded that all three radiation components are generated by the ordered (nonthermal) motion of electrons.

523.72:523.746 1981

A Method of Analysis of the Directivity of Solar Radio Emission from Sunspots—T. Takakura. (*Nature* (*London*), vol. 171, p. 445; March 7, 1953.)

523.72:531.758:523.745 1982

An Estimate of the Density and Motion of Solar Material from Observed Characteristics of Solar Radio Outbursts—H. K. Sen. (*Aust. Jour. Phys.*, vol. 6, pp. 67–72; March 1953.) Theory previously given by Feinstein and Sen (104 of 1952) is extended and applied to the observations made in Australia by Wild and McCready (1629 of 1951) and by Wild (1630, 2169 and 2170 of 1951) of the spectra of solar RF outbursts in the range 70–130 mc. The dispersion equation, involving the velocity of solar material erupting into a static corona and the temperatures and densities of the material and of the corona, enables the velocity of the material to be estimated as about 500 km/s, with a particle density of about 10^8/cm³.

523.74/.75:[550.38+551.510.5 1983

Solar and Terrestrial Relationships—D. H. McIntosh. (*Met. Mag.* (*London*), vol. 82, pp. 11–15; Jan. 1953.) The general problem of the relation between solar activity on the one hand and meteorological, geomagnetic and ionospheric phenomena on the other is reviewed, with special reference to an account by Scherhag (*Wetterkarte*, No. 74, March 14, 1952) of events which he interprets as being related to an intense solar eruption on Feb. 24, 1952. No such flare has been reported. It is concluded that no clear evidence has yet been given of a direct relation between solar phenomena and meteorological effects.

523.746"1952.10/.12" 1984

Provisional Sunspot-Numbers for October to December, 1952—M. Waldmeier. (*Jour. Geophys. Res.*, vol. 58, p. 112; March 1953.)

538.122:523 1985

The Origin of Magnetic Fields in Moving Plasma—L. Biermann. (*Ann. Phys., Lpz.*, vol. 10, pp. 413–417; July 15, 1952.) Discussion indicates that diffusion effects produce sufficiently large emf's in rotating masses of ionized gas to account for the existence of solar and stellar magnetic fields.

550.372:621.396.11 1986

Effect of a Large Dielectric Constant on Ground-Wave Propagation—Wait and Campbell. (See 2113.)

550.38"1952.10/.12" 1987

Cheltenham [Maryland] Three-Hour-Range, Indexes *K* for October to December, 1952—R. R. Bodle. (*Jour. Geophys. Res.*, vol. 58, p. 112; March 1953.)

550.385"1952.07/.12" 1988

Principal Magnetic Storms [July–Dec. 1952]—(*Jour. Geophys. Res.*, vol. 58, pp. 113–115; March 1953.)

550.386 1989

International Data on Magnetic Disturbances, Third Quarter, 1952—J. Bartels and J. Veldkamp. (*Jour. Geophys. Res.*, vol. 58, pp. 109–111; March 1953.)

551.510.3:535.325 1990

The Constants in the Equation for Atmospheric Refractive Index at Radio Frequencies—E. K. Smith, Jr. and S. Weintraub. (*Jour. Res. Nat. Bur. Stand.*, vol. 50, pp. 39–41; Jan. 1953.) The formula for scaled-up refractivity N for air is $N=(K_1/T)[p+K^1_2(e/T)]$, where K_1, K^1_2 are constants. The value proposed for K_1 is derived from recent reliable microwave and optical measurements of the refractive index of dry air by Barrell (*Jour. Opt. Soc. Amer.*, vol. 41, pp. 295–299, 1951, Birnbaum *et al.* (1426 of 1951) and Essen and Froome (1371 of 1952), after adjustment to allow for an average amount of CO_2 in the composition of the air. K^1_2 is derived from two other constants obtained from measurements of the dielectric constant of water vapour by Birnbaum and Chatterjee [2269 of 1952 (please note additional author)]. A table of values of the constants used by various workers from 1933 to 1952 is given, with comments on discrepancies.

551.510.52 1991

The Geographical and Height Distribution of the Gradient of Refractive Index—B. R. Bean. (Proc. I.R.E., vol. 14, pp. 549–550; May 1953.) Charts are presented of the February and August distributions of the effective earth's radius factor over the U.S.A., and of the vertical distribution of refractive-index gradient for warm, temperate and cold climates.

551.510.535 1992

Long-Delay Ionospheric Echoes at 150 kc—J. J. Gibbons, R. L. Schrag and A. H. Waynick. (*Nature* (*London*), vol. 171, pp. 444–445; March 7, 1953.) A report of results obtained at Pennsylvania State College Ionosphere Research Laboratory, using a vertical-incidence pulse transmitter with 2-MW peak power. Recordings with a 5-min exposure were made every half hour from 2000 to 0400 LMST of the signals received on an antenna system suitable for receiving circularly polarized waves. Three records, obtained on November 9, 1952, showed an isolated echo corresponding to a reflection height of about 900 km. The echo only appeared on the records of the extraordinary wave. Similar records were obtained on two subsequent nights. Possible explanations of these long-delay echoes are considered; they may be due to reflections from a region above the height of maximum ionization of the F_2 region.

551.510.535 1993

On the Comparative Increases of the F_1 and F_2 Ionizations from Sunspot Minimum to Sunspot maximum—M. Ghosh. (*Jour. Geophys. Res.*, vol. 58, pp. 41–51; March 1953.) The proportionate increase of ionization, from the epoch of sunspot minimum to sunspot maximum, is not the same for the F_1 as for the F_2 layer, being of the order of 1.6 for the former and 3.0 for the latter. Assuming (*a*) the F region to be due to ionization of atomic oxygen, (*b*) an increase in temperature with height above the F_1 maximum, (*c*) an effective recombination coefficient decreasing rapidly with height, calculations are made for three different ionosphere models, the results being in each case in satisfactory agreement with observations.

551.510.535:523.78 1994

Effect of the Solar Eclipse of 25th February 1952 on the Ionosphere E Layer in Equatorial Africa—S. Estrabaud. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 833–835; Feb. 23, 1952.) A report and discussion of observations of E-layer critical frequencies at Bangui. The critical frequency of the E_2 layer decreased suddenly at the commencement of the eclipse and was soon masked by that of the E_1 layer. A graph of the observed values of f_0/E shows that the effects of the eclipse were certainly perceptible some 5 to 20 minutes before the first contact. An explanation of the observed effects is based on the assumption of two solar regions more active than the rest, one in the corona near the western edge of the disk, the other near the eastern edge.

551.510.535:[537.29+538.69 1995

Electromagnetic Effects in the Ionosphere: Effects of Crossed Constant-Magnetic and Alternating-Electric Fields—R. Jancel and T. Kahan. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 788–790; Feb. 23, 1953.) The effect of the crossed fields on the distribution of the velocities of the free electrons in an ionized gas is studied by means of the general equation of Boltzmann, collisions between electrons being assumed negligible. The result obtained includes, as particular cases, those of Druyvestyn (1930 Abstracts, p. 350 and 1700 of 1935), Chapman and Cowling (The Mathematical Theory of Nonuniform Gases) and Margenau (3246 of 1946).

551.510.535:551.55:621.396.9 1996

A Study of Winds in the Ionosphere by Radio Methods—J. H. Chapman. (*Canad. Jour. Phys.*, vol. 31, pp. 120–131; Jan. 1953.) Report of an investigation made using Mitra's method (96 of 1950). The measurements were made in Ottawa, and later in Montreal, and covered a period of a year. At a nominal height of 110 km (E layer) the speed of the wind showed a mean daily variation of approximately half-day period and about 40 m/s amplitude. This variation can be explained on the basis of tidal oscillations in the atmosphere. A variation corresponding to lunar atmospheric tides has also been detected at this level. The winds in the F region appear to increase in speed with increase of magnetic activity; this effect is not observed in the E region except during severe ionospheric disturbances.

551.510.535:551.594.12 1997

A Procedure for the Determination of the Vertical Distribution of the Electron Density in the Ionosphere—L. A. Manning. (*Jour. Geophys. Res.*, vol. 58, pp. 117–118; March 1953.) Comment on 409 of February (Kelso).

551.510.535:621.3.087.4 1998

An Ionosphere Recorder for Low Frequencies—J. C. Blair, J. N. Brown and J. M. Watts. (*Jour. Geophys. Res.*, vol. 58, pp. 99–107; March 1953.) Based on HF recorder design, this beat-frequency instrument covers the range 50 kc–1 mc and uses a single-loop antenna about 350 ft long by 150 ft high. Sample records are reproduced.

551.594.5 1999

Radio Measurements and Auroral Electron Densities—P. A. Forsyth. (*Jour. Geophys. Res.*, vol. 58, pp. 53–66; March 1953.) Simultaneous recordings of 56- and 106.5-mc echoes from auroral clouds were made at Saskatoon. A large number of echoes at 56 mc were found for which there were no corresponding echoes at 106.5 mc, and the echo amplitude ratio for the two frequencies varied widely. Critical reflections from small volumes of intense ionization seem the most probable explanation of these observations, where intense auroral displays are concerned. This would involve electron densities of the order of 10^8/cm^3 in small volumes of the aurora for periods of time short in comparison with the duration of the display.

551.594.5+523.5+551.510.535:621.396.9 2000

Long-Duration Echoes from Aurora, Meteors, and Ionospheric Back-Scatter—D. W. R. McKinley and P. M. Millman. (*Canad. Jour. Phys.*, vol. 31, pp. 171–181; Feb. 1953.) A report of radar echoes of unusual types observed in the course of the Ottawa meteor-research program. 20-μs pulses with a peak power of 200–400 kW were radiated on a frequency of 33 mc. Some of the echoes were correlated with visual observations of aurora. The extremely long duration (9 min to over 30 min) of some meteor echoes was probably caused by abnormal ionospheric conditions. Weak semipermanent echoes (duration up to or over 1 hour) are attributed to reflection at vertical or near-vertical incidence from irregularities in the lower ionosphere, possibly the same as those responsible for the long-range propagation reported by Bailey *et al.* (2581 of 1952).

LOCATION AND AIDS TO NAVIGATION

621.396.9 2001

High-Quality Radar Picture Display Unit—R. T. Petruzzelli. (*Tele-Tech*, vol. 12, pp. 54–57, 102; Jan. 1953.) A description is given, with block diagrams, of equipment providing television pictures of the PPI display of airport-surveillance radar, Type ASR-2, together with mapping and df information. Two independent channels, each terminating in its own 12-in. display unit, enable different areas round the airport to be monitored. A 30-in. display for plotting purposes is also provided. The conversion from the PPI display to standard 525-line 30-frame television scan is effected by means of the graphechon storage tube [1539 of May—Dyall *et al.*]. The display control consoles include all controls for the radar set normally found on the original radar indicator. The plotting display console has no control over the radar set and can merely switch its mixer amplifier to accept the signals from either of the two channels.

621.396.9 2002

A Special Type of Pulse Modulation for the Transmission of Radar Pictures (Sampling)—E. Roessler. (*Fernmeldetech. Z.*, vol. 6, pp. 78–79; Feb. 1953.) The principles of a sampling system for bandwidth reduction are described. See 1921 (Otto) and 3430 (McLucas) of 1952. It is proposed to introduce, in German, the term "Durchmusterung" for this sampling process.

621.396.9:[551.578+551.594.22 2003

Frontal Precipitation and Lightning Observed by Radar—J. S. Marshall. (*Canad. Jour. Phys.*, vol. 31, pp. 194–203; Feb. 1953.) Patterns obtained by radar methods at wavelengths of 3.2 cm and 10.7 cm are shown and discussed. Observations in vertical section were made by scanning at angles of elevation up to 30°.

621.396.9:551.578.1 2004

Radar Observations of Rain from Nonfreezing Clouds—R. S. Styles and F. W. Campbell. (*Aust. Jour. Phys.*, vol. 6, pp. 73–85; March 1953.) A detailed account, with photographs of typical displays, of results obtained with airborne radar equipment.

621.396.9:621.396.65.029.64 2005

Remote Display of Radar Pictures—R. F. Hansford and G. J. Dixon. (*Wireless World*, vol. 59, pp. 218–222; May 1953.) Description of the Decca Radar Link, Type 2, designed for transmission of information from a radar site to a distant point where it can again be displayed on a radar indicator without loss of detail. The wavelength used is about 9 cm, and the transmitter power 0.5 W. Negative a.m. is used for the video signals and positive modulation on a time basis for the synchronizing pulses and bearing information. The transmitter RF unit is housed in a watertight box mounted behind the paraboloid antenna on an 80-ft mast. The transmitter tube is a v.m. Heil-type tube. A similar antenna arrangement is used with the receiver, which has a reflex-klystron local oscillator. Full test facilities are included. Reliable transmission is provided over a distance up to 20–30 nautical miles. Repeater stations may be interpolated in the link.

621.396.9:621.396.67 2006

Moulded Plastic Radar Scanners and Stressed Components—(See 1911.)

621.396.9:778.5 2007

Cinematographic Recording of Panoramic Radar Images—M. J. de Cadenet. (*Ann. Télécommun.*, vol. 8, pp. 19–27; Jan. 1953.) A method developed by the C.N.E.T. in France for recording television images is adapted. A detail description is given of camera equipment with servomechanism for synchronizing camera operation with antenna rotation.

621.396.93.088 2008

More on Direction Finders—H. G. Hopkins. (Proc. I.R.E., vol. 41, pp. 548–549; May 1953.) Comment on 142 of 1952 (Hansel). It is recommended that the term 'unwanted' should be reserved for the component which introduces errors when the direction finder is operated on a uniform unobstructed site. The proposed distinction between primary and secondary instrumental polarization errors is also criticized.

621.396.932/.933 2009

Decca Charts—H. C. Freiesleben. (*Telefunken Ztg*, vol. 26, pp. 49–53; Jan. 1953.) A detailed description is given of the method of computing Decca curves in which the differences of the distances of a number of points on the earth's surface from the transmitters are first calculated, the points corresponding to round values of the differences being then found by interpolation.

621.396.933 2010

Radio Navigation and Air Traffic Safety Measures—O. J. Selis. (*Tijdschr. ned. Radiogenoot.*, vol. 18, pp. 59–87; March 1953.) A survey of present methods and trends both for en-route navigation and for approach and landing, with special reference to installations in Holland.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 2011

Methods of Obtaining High Vacua by Ionization—H. Schwarz. (*Le Vide*, vol. 7, pp. 1262–1266; Nov. 1952.) Analysis shows that no true pumping effect can be obtained with a simple d.c. discharge. In the system described a ring-shaped anode has a potential of 2kV or more and a second annular electrode beyond it is at cathode potential. Four outlet tubes fitted with auxiliary electrodes are spaced around the anode. A weak magnetic-field imparts a helical motion to the electrons beyond the anode. Pressures lower than 5×10^{-6} mm Hg are obtained.

535.212:546.482.21 2012

Study of the Infrared Photoluminescence of Copper-Activated Cadmium Sulphide—E. Grillot and P. Guintini. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 802–804; Feb. 23, 1953.) The photoluminescence of CdS-Cu comprises a band in the infrared between 0.7 and 1.4 μ, with a maximum at 1.02 μ.

535.215:546.817.221 2013

The Photovoltaic Effect in Natural Lead Sulphide—R. Lawrance. (*Aust. Jour. Phys.*, vol. 6, pp. 124–125; March 1953.) Correction to paper abstracted in 1930 of 1952.

535.215.2:546.28 2014

Long-Wavelength Infrared Photoconductivity of Silicon at Low Temperatures—B. V. Rollin and E. L. Simmons. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 162–168; March 1, 1953.)

535.215.3:538.639 2015

Theory of the Photomagnetoelectric Effect—P. Aigrain and H. Bulliard. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 595–596; Feb. 9, 1953.) An approximate theory is developed which assumes the charge carriers of the two different types in semiconductors to have equal mobilities and which gives values of the photomagneto-electric potential difference in good agreement with the experimental results reported by Kikoin and Noskow (Abstracts, p. 507; 1934).

535.215.3:538.639:546.289:548.55 2016

Experimental Results on the Photomagnetoelectric Effect—P. Aigrain and H. Bulliard. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 672–674; Feb. 16, 1953.) Measurements were made on small plates, 0.1 to 1 mm thick, cut from single crystals of Ge of various degrees of purity, the illuminated face being in general well polished chemically. Various methods of measurement are described. The results obtained confirm the theory previously given (2015 above).

535.343+535.32 2017

Inter-relation between Optical Constants for Lead Telluride and Silicon—T. S. Moss. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 141–144; Feb. 1, 1953.)

535.343 2018

The Optical Constants of Lead Sulphide, Lead Selenide and Lead Telluride in the 0.5–3-μ Region of the Spectrum—D. G. Avery. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 134–140; Feb. 1, 1953.)

535.343:546.23 2019

The Optical Absorption and Reflection of Amorphous and Hexagonal Selenium—J. Stuke. (*Z. Phys.*, vol. 134, pp. 194–207; Jan. 20, 1953.) Measurements were made on thin layers, using wavelengths between 300 and 1200 mμ, at temperatures in the range −180° C to +100° C. At the transformation from the amorphous to the hexagonal form the absorption constant increases over the whole of the wave-range investigated. For hexagonal Se a strong absorption band is observed at 2.2 eV. The absorption and reflection measurements show good agreement.

535.343+537.312.5:546.817.221 2020

Absorption Spectra of Lead Sulphide at Different Temperatures—W. Paul and R. V. Jones. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 194–200; March 1, 1953.)

535.37 2021

Induced Conductivity in Luminescent Powders: Part 2—A.C. Impedance Measurements—H. Kallmann, B. Kramer and A. Perlmutter. (*Phys. Rev.*, vol. 89, pp. 700–707; Feb. 15, 1953.) A.C. impedance changes induced by ultraviolet, infrared, X-ray and γ-ray irradiation of (Zn:Cd)S powders agree with the results previously obtained with d.c. (Kallmann and Kramer, 3438 of 1952.)

535.37:546.472.21 2022

The Influence on the Luminescence of ZnS Phosphors of Lattice Defects caused by Pounding—I. Broser and W. Reichardt. (*Z. Phys.*, vol. 134, pp. 222–224; Jan. 20, 1953.) An investigation is made of the reduction of luminescence of ZnS-Cu phosphors as the fineness of the material increases. Formulas derived are discussed in relation to theories of radiationless transitions.

535.37:546.472.21 2023

Optical Measurements on Electroluminescent Zinc Sulfide—J. F. Waymouth. (*Jour. electrochem. Soc.*, vol. 100, pp. 81–84; Feb. 1953.) Results of measurements of absorption and emission spectra are reported.

535.37:546.472.21 2024

On the Nature of Fluorescent Centers and Traps in Zinc Sulfide—H. A. Klasens. (*Jour. electrochem. Soc.*, vol. 100, pp. 72–80; Feb. 1953.) The impurity levels in the energy diagram of a ZnS phosphor are considered to be localized S^{2-}-levels lifted above the filled S^{2-}-band due to the presence of monovalent positive or trivalent negative activator ions in the lattice. Electron traps are formed similarly by the replacement of S ions by monovalent negative ions or of Zn^{2+} ions by trivalent positive ions. The emission characteristics of ZnS phosphors seem to be more a property of the lattice than of the activator.

537.224 2025

Plastic Electrets—H. H. Wieder and S. Kaufman. (*Jour. Appl. Phys.*, vol. 24, pp. 156–161; Feb. 1953.) Previously proposed explanations of the electret mechanism are discussed and measurements on specimens made of plexiglas, lucite and nylon are reported. The experimental results support the view that there exist two decaying polarizations of opposite sense due to ionic migration (*a*) within the dielectric and (*b*) across the electrode/dielectric interface (see Gross, 4055 of 1944.)

537.226 2026

Variation with Time of the Dielectric Properties of Ferroelectric Ceramics—D. M. Kazarnovski. (*Zh. tekh. Fiz.*, vol. 22, pp. 553–558; April 1952.) A report of observations covering a period of three years. It appears that the dielectric constant of ferroelectric ceramics decreases with the time; methods for stabilizing it are discussed.

537.226.2:546.212-16 2027

Anistropy of the Dielectric Constant of Ice—F. Humbel, F. Jona and P. Scherrer. (*Helv. Phys. Acta*, vol. 26, pp. 17–32; Feb. 15, 1953. In German.) Dielectric constant ε and loss angle δ were measured as a function of temperature and frequency for square plates cut from large single crystals. The difference between the values of ε for directions parallel and perpendicular to the *c*-axis decreases with temperature decrease. The variation of ε and δ with frequency is similar to that found for polycrystalline ice.

537.311.33 2028

Theory of Mixed Semiconductors—O. Madelung and H. Welker. (*Z. angew. Phys.*, vol. 5, pp. 12–14; Jan. 1953.) For simple *n*- or *p*-type semiconductors the carrier concentrations and mobilities can be easily found from the measured values of conductivity and Hall coefficient. For mixed semiconductors further measurements are required; the thermoelectric power and the variation of the resistance with the strength of an applied magnetic field are suitable properties for this purpose. Formulas relating the various parameters are discussed. The Hall coefficient depends on the strength of the magnetic field and may change its sigu at high field strength.

537.311.33 2029

A Conductive Ceramic for Microwave Applications—(*Eng.* (*London*), vol. 195, p. 363; March 6, 1953.) "Caslode" is the name given to a semiconductor with a dielectric constant of about 20 and a loss angle between 10° and 20° at a wavelength of 3.2 cm, suitable for use without water-cooling as a dummy laod in waveguides.

537.311.33 2030

Semiconductor Properties of the Systems K-Sb, Cs-Sb, K-In and Cs-In—R. Suhrmann and C. Kangro. (*Naturwissenschaften*, vol. 40, pp. 137–138; Feb. 1953.) Graphs are given which show (*a*) log χ for K-Sb compounds with varying proportions of K from 0 to 100%, (*b*) log χ plotted against 1/T for values of 1/T in the range $3–9\times10^{-3}$ for the compounds Cs_3Sb and K_3Sb, χ being the specific conductivity and T the absolute temperature. For intrinsic semiconduction, the energy levels in K_3Sb, KSb and Cs_3Sb are respectively 0.79, 0.88 and 0.56 eV. For lattice-defect semiconduction, the levels are 0.23 and 0.16 eV respectively for K_3Sb and Cs_3Sb. For K-In and Cs-In compounds also, the temperature coefficient of resistance is negative; for Cs_2In_3 the resistance drops by a factor of about 10^4 between 83° and 287°K, the corresponding drop for K_3In being only from 82 to 37.7Ω.

537.311.33:546.27.03 2031

Study of the Properties of Boron—J. Lagrenaudie. (*Jour. Phys. Radium*, vol. 14, pp. 14–18; Jan. 1953.) Continuation of work noted in 1696 of June. Conductivity measurements at temperatures up to 700° C yield a value of about 1.28 eV for the intrinsic energy. The material should be useful for thermistors. Thin layers are transparent in the infrared, with an absorption electric threshold at a wavelength near 1μ. Only very weak rectifying effects were observed; measurements on single crystals are required before conclusions can be reached about the practical possibilities of the material for rectifiers.

537.311.33:546.289 2032

A Thermoelectric Study of the Electrical Forming of Germanium Rectifiers—M. Kikuchi and T. Onishi. (*Jour. Appl. Phys.*, vol. 24, pp. 162–166; Feb. 1953.) Experimental methods described by Granville and Hogarth (161 of 1952) were applied. Remarkable improvements in the I/V characteristics were obtained by applying appropriate alternating forming voltages. Using *n*-type crystals, the thermoelectric current observed on the etched surface was converted by the forming from *n*-type to *p*-type. The dependence of the thermo-emf on the contact pressure is greatly increased by the forming. The results indicate that a substance with relatively high resistivity and *p*-type thermo-emf is formed between the point contact and the Ge surface.

537.311.33:546.289 2033

Plastic Deformability of Germanium at Relatively High Temperatures—L. Graf, H. R. Lacour and K. Seiler. (*Z. Metallkde*, vol. 44, pp. 113–114; March 1953.) Experiments on single crystals of purity >99.999% are reported. At temperatures >600° C plastic deformation without time lag was obtained. The appearance of slip lines indicates that an actual lattice translation occurs. Variation of resistivity caused by heat treatment is greatly reduced if the specimens are simultaneously subjected to plastic deformation. See also 1403 of May (Gallagher).

537.311.33:546.289 2034

Pressure-Welded *p-n* Junctions in Germanium—R. G. Shulman and D. M. Van Winkle. (*Jour. Appl. Phys.*, vol. 24, p. 224; Feb. 1953.) A mechanically strong weld is obtained by applying local pressure at the surfaces to be joined. Pressures of about 100 kg/cm² are obtained with moderate forces by making the contact area extremely small; this is achieved by making one contact surface optically flat and the other of controlled roughness. Characteristics of junctions prepared by this process are shown; their HF response is better than that of units prepared by drawing or diffusion.

537.311.33:546.289:621.396.822 2035

Shot Noise in Germanium Filaments—R. H. Mattson and A. van der Ziel. (*Jour. Appl. Phys.*, vol. 24, p. 222; Feb. 1953.) The correctness of the noise formula previously derived (Herzog and van der Ziel, 1936 of 1952) is verified by further measurements of greater accuracy over an extended frequency range.

537.311.33:546.817.221 2036

Transistor Action and Related Phenomena in Lead Sulphide Specimens from Various Sources—C. A. Hogarth. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 216–220; March 1, 1953.) Examination of a large number of specimens of PbS has indicated a maximum free-carrier concentration at 290°K of $2\times10^{17}/cm^3$ if transistor action is to be found. This leads to a value of 0.65 eV for the width of the forbidden band at 290°K. Transistor action was only observed in *p*-type specimens and was always associated with good rectification and a strong photovoltaic effect.

537.311.33:546.817.221 2037

The Electronic Band Structure of PbS—D. G. Bell, D. M. Hum, L. Pincherle, D. W. Sciama and P. M. Woodward. (*Proc. Roy. Soc. A* (*London*), vol. 217, pp. 71–91; March 24, 1953.) The cellular method is used to determine electronic wave functions of the Bloch type for PbS.

537.311.33:621.314.632 2038

The Theory of Solid Rectifiers—A. I. Gubanov. (*Zh. tekh. Fiz.*, vol. 22, pp. 381–393; March 1952.) A theory is proposed based on the assumption that the barrier layer is a thin semiconducting layer with a conductivity of the type oposite to that of the main body of the rectifier. From a solution of the Poisson's and diffusion equations the current/voltage characteristic of the rectifier is determined. A numerical example is given for the case of a Cu_2O rectifier; the theoretical curve for the dependence of the rectifier resistance on voltage shows good agreement with experimental curves.

537.311.33:621.396.822 2039

Noise in Semiconductors at Very Low Frequencies—B. V. Rollin and I. M. Templeton. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 259–261; March 1, 1953.) Previous measurements by Kronenberger (2879 of 1951) showed that down to frequencies of about 0.2 cps the mean-square noise voltage per unit bandwidth is inversely proportional to the freauency f. A method is described by which measurements on pyrolytic carbon resistors have been made at much lower frequencies. The noise is recorded on slow-running magnetic tape, which is then joined to form a closed loop. The loop is run through the pickup head at normal speed and the noise spectrum analyzed with a standard type of AF wave analyzer. By varying the recording speed, measurements were made for frequencies from 10^{-4} to 10 cps. The results obtained show that even at the lowest frequency the $1/f$ law is still obeyed fairly closely. A special arrangement is described for obtaining the amplification necessary for recording the noise voltages at the very low frequencies used.

537.311.33:621.396.822 2040

Shot Noise in Semiconductors—A. van der Ziel. (*Jour. Appl. Phys.*, vol. 24, pp. 222–223; Feb. 1953.) A calculation is made of the shot noise generated in a semiconductor by dc, assuming that the numbers of free electrons and holes fluctuate independently, and the average drift path is short compared with the length of the specimen. An examination is made of the special cases of electron conductors, mixed conductors, intrinsic semiconductors in which electrons and holes disappear by surface trapping, and intrinsic semiconductors in which electrons and holes disappear only by recombination.

538.221 2041

Magnetic After-Effects associated with Initial Permeability and Barkhausen Jumps—R. Feldtkeller and G. Sorger. (*Arch. elekt. Übertragung*, vol. 7, pp. 79–87; Feb. 1953.) A discussion of the Jordan, Richter and Ewing effects observed in relaxation processes in ferromagnetic materials; the first two are associated with the initial permeability and the last is associated with the Barkhausen jumps. The Richter and Ewing effects both depend on the diffusion of foreign atoms.

538.221 2042

Applications and Properties of Ferroxcube—(*Electronic Applic. Bull.*, vol. 13, p. 80; May 1952.) Correction to paper noted in 750 of March.

538.221 2043

Ferrites—A. Fairweather, F. F. Roberts and A. J. E. Welch. (*Rep. Progr. Phys.*, vol. 15, pp. 142–172; 1952.) A description is given of the physical and chemical structure of ferrites, and of their preparation, with an outline of Néel's theory of ferromagnetism and of the Kramers-Anderson theory of super-exchange spin coupling. Possible mechanisms which may account for the observed rather steep fall of permeability with increase of frequency, and for the associated energy absorption, are discussed. The dielectric properties of ferrites, of interest in HF applications, are considered. The observed very strong dispersion and absorption may be interpreted in terms of the inhomogeneous microphysical structure of the sintered material, which is known to be sensitive to the oxygen content and heat treatment. The very-low-frequency dielectric absorption may be due to adsorbed moisture. The semiconducting properties are discussed briefly. Over 150 references.

538.639:538.245 2044

On the Temperature Dependency of Magneto-resistance Effect of Iron Single Crystal—Y. Gondo and Z. Funatogawa. (*Jour. Phys. Soc.* (*Japan*), vol. 7, pp. 41–43; Jan./Feb. 1952.

538.652:548.55 2045

Measurement of Magnetostriction in Single Crystals—R. M. Bozorth and R. W. Hamming. (*Phys. Rev.*, vol. 89, pp. 865–869; Feb. 15, 1953.) A simplified procedure is given for determining the five magnetostriction constants of a ferromagnetic single crystal of the cubic type.

539.153:537.311.1 2046

The Band Structure of Metals—G. V. Raynor. (*Rep. Progr. Phys.*, vol. 15, pp. 173–248; 1952.) The development of the electron theory of metals is briefly reviewed and the electronic structures of a selection of monovalent and polyvalent metals are surveyed, Na and Cu being discussed in detail. Difficulties arising in connection with the calculation of electronic interactions, particularly in the case of divalent and trivalent metals, are discussed. The nature of transitional metals is considered in detail. The possibility of extending the theory to alloys is discussed, particularly for alloys based on Cu and on Ni. Over 80 references.

546.321.85-842:548.0[537+539.32 2047

The Properties of KH_2PO_4 below the Curie Point—H. M. Barkla and D. M Finlayson (*Phil Mag.*, vol. 44, pp. 109–130; Feb. 1953.) Report of an investigation of the dielectric, piezoelectric and elastic properties of KH_2PO_4 crystals at temperatures down to 20°K. A second transition occurs at a temperature about 60° below the Curie point (122°K), and marks the onset of a steep rise of coercive field with decreasing temperature. A qualitative domain theory is presented to account for the main features of the ferroelectric state in KH_2PO_4.

546.321.85-842:548.73 2048

X-Ray Analysis of the Ferroelectric Transition in KH_2PO_4—B. C. Frazer and R. Pepinsky. (*Acta Cryst.*, *Camb.*, vol. 6, part 3, pp. 273–285; March 10, 1953.) Report of detailed investigation of the tetragonal structure at 4° above the transition temperature of 122°K, and of the orthorhombic structure 6° below this temperature.

546.821:538.632 2049

The Hall Effect in Titanium—G. W. Scovil. (*Jour. Appl. Phys.*, vol. 24, pp. 226–227; Feb. 1953.) Preliminary report of measurements made by a dc method at a temperature of about 100° C.

547.476.3:537.226 2050

The Effect of Electric Field on the Domain Structures in Rochelle Salt—M. Marutake. (*Jour. Phys. Soc.* (*Japan*), vol. 7, pp. 25–29; Jan./Feb. 1952.)

549.212:537 2051

The Electrical Properties of Graphite—G. H. Kinchin. (*Proc. Roy. Soc. A*, vol. 217, pp. 9–26; March 24, 1953.) Report of measurements made, over a wide range of temperatures, of the Hall coefficient and resistivity of a range of multicrystal graphites with different crystal sizes and of a single crystal of Travancore graphite.

621.315.61:621.793:669.21/.23 2052

The Production and Testing of Precious-Metal Coatings on Insulating Materials—A. Keil and G. Offner. (*Fernmeldetech. Z.*, vol. 6, pp. 73–77; Feb. 1953.) Methods of metallizing the surfaces of plastics and ceramics are outlined and suitable methods of assessing the properties of such coatings, particularly as regards resistivity, adhesion and soldering, are suggested.

621.315.612.4:546.431.824-31 2053

Transition Energy and Volume Change at Three Transitions in Barium Titanate—G. Shirane and A. Takeda. (*Jour. Phys. Soc.* (*Japan*), vol. 7, pp. 1–4; Jan./Feb. 1952.)

621.315.616:547-128† 2054

Silicones—and Insulation—(*Elect. Times*, vol. 123, pp. 47–49; Jan. 8, 1953.) Illustrated note of applications of silicone rubber as sheathing and insulating tape.

621.315.616.1 2055

Electrical Properties of Rubber—J. Grannier. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 786–788; Feb. 23, 1953.) Experiments are described which show that the dc conductance and the dielectric constant and loss angle at 1 kc of rubber decrease considerably when the material is subjected to increasing compressional, tensile, or shearing forces with resulting distortion. When the material is compressed in a cavity which prevents distortion, little or no change is observed in the dielectric constant and loss angle. An explanation of these results is suggested.

778.37 2056

Isotransport Camera for 100,000 Frames per Second—C. D. Miller and A. Scharf. (*Jour. Soc. Mot. Pict. Telev. Engs.*, vol. 60, pp. 130–144; Feb. 1953.) Description of equipment developed by the Battelle Memorial Institute and available for loan to other research organizations.

MATHEMATICS

621.392.5:538.652:681.142 2057

Magnetostrictive Sonic Delay Line—H. Epstein and O. Stram. (*Rev. Sci. Instr.*, vol. 24, pp. 231–232; March 1953.) Description of a recirculating storage unit using a thin-wall Ni tube as the magnetostrictive element, with suitable transducer coils enclosed in ferrite cups whose length is adjusted to correspond to the half wavelength of a 600-kc vibration in the Ni tube. Experimental delay lines giving delays of 100μs and 800 μs have been constructed. Operational characteristics are listed.

681.142 2058

High-Speed Product Integrator—A. B. Macnee. (*Rev. Sci. Instr.*, vol. 24, pp. 207–211; March 1953.) A method of using a high-speed analogue computer to evaluate product integrals is described.

681.142 2059

Resistance Network Analog Computer—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 37, pp. 19–21; Feb. 1953.) Description of a computer of the type developed by Liebmann (1954 of 1950 and 2839 of 1952).

681.142 2060

The Electronic Discrete Variable Computer—S. E. Gluck. (*Elec. Eng.*, *N. Y.*, vol. 72, pp. 159–162; Feb. 1953.) Outline description of the EDVAC, a binary computer having a mercury-delay-line storage system of large capacity. See also 434 (Goodman) and 1357 (Gray) of 1952.

681.142:538:221 2061

Ferrites speed Digital Computers—D. R. Brown and E. Albers-Shoenberg. (*Electronics*, vol. 26, pp. 146–149; April 1953.) The use of ferrite toroids in storage units of the matrix type [2258 of 1952] (Papian) leads to increased speed and reliability of operation. Differences in the hysteresis curves required for storage and for switching purposes are discussed. A pulse method of testing the toroids is described. A particular Mg ferrite (MF-1118) used at the M.I.T. has a saturation flux density of 2000 gauss and a coercivity of 1.5 oersted.

MEASUREMENTS AND TEST GEAR

535.32:538.56.029.64 2062

The Refractive Indexes of Water Vapour, Air, Oxygen, Nitrogen, Hydrogen, Deuterium and Helium—L. Essen. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 189–193; March 1953.) The method described by Essen and Froome (1707 of 1951) was used in measurements at 9.2 kmc.

The results for water vapour, air, O_2 and N_2 agree with the values previously found at 24 kmc. The values for H_2 and D_2 are respectively 2% and 3% higher than the calculated values of Ishiguro *et al.* (*Proc. Phys. Soc.* (*London*), vol. 65, pp. 178–187; March 1, 1952.)

621.3.018.41(083.74) 2063

The Microwave Frequency Standard—L. J. Rueger and A. E. Wilson. (*Radio & Telev. News, Radio-Electronic Eng. Sec.*, vol. 49, pp. 5–7, 41; March 1953.) An account is given of standardized equipment at the National Bureau of Standards for providing a calibration service for frequencies from 300 mc to 40 kmc, and of developmental equipment to extend the range to 75 kmc. The standard microwave frequencies are derived directly from one of the stable 100-kc crystal oscillators maintained by the N.B.S. Harmonic frequencies at 10-mc intervals are provided up to 5 kmc, at 50-mc intervals to 25 kmc, and at 250-mc intervals to 40 kmc, the gaps being bridged by means of a precision variable-frequency oscillator. Frequency meters are calibrated under conditions as close as possible to those of normal use. Details are given of the methods adopted.

621.3.018.41(083.741.):[621.314.7+621.396.611.21 2064

Precision Transistor Oscillator—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 37, pp. 17–19; Feb. 1953.) A frequency standard contained in a tube of length 7 in. and diameter 1¾ in. is described. It comprises an oscillator using a Type-2517 junction transistor, a high-precision 100-kc GT-cut quartz-crystal unit [see Griffin, 1083 of April] and a long-life Hg cell. The transistor in an earthed-emitter circuit produces an output of 0.8V across a tuned circuit connected to the collector electrode. The crystal driving current is taken from the junction of two capacitors forming an attenuator between the collector and earth. Performance is comparable to that of a standard valve oscillator. (For another account, see *Electronics*, vol. 26, pp. 206, 214; May 1953.)

621.317.32.087.4 2065

The Suitability of the "Fixed Level" Recording Method for Propagation Research—G. Boré and W. Rappaport. (*Fernmeldetech. Z.*, vol. 6, pp. 33–36; Jan. 1953.) The design of an u.s.w. field-strength recorder is outlined. Its principle is that described by Ferrari (*A.E.G. Mitt.*, vol. 41, pp. 299–302; Nov./Dec. 1951), a pulse being generated whenever the integral of the measured quantity over a specified time reaches a fixed level. The average interval between pulses is inversely proportional to the mean amplitude of the measured quantity. A method of evaluating the maximum error is illustrated. By means of a subsidiary relay-operated circuit, the number and duration of short-period fluctuations above and below a fixed level can be recorded.

621.317.335.3:537.226.2/.3 2066

A Standing-Wave Method for Measuring Electromagnetic Absorption in Polar Liquids at Frequencies of the Order 3×10^9 cps—V. I. Little. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 175–184; March 1, 1953.) Description of a simple method, with typical results for the absorption coefficient and dielectric constant of water and of dioxan with small admixtures of water and of ethyl alcohol.

621.317.335.3.029.64 2067

New Method of Measurement of the Complex Dielectric Constant of Solids and Liquids at Centimeter Wavelengths—J. Le Bot and S. Le Montagner. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 469–471; Feb. 2, 1953.) A Method is described which requires only very small quantities of the material under test. Solid samples in the form of rods, or liquid samples contained in thin-walled glass tubes are arranged as shunt elements transverse to the larger sides of a rectangular waveguide. Using a short-circuited waveguide section and fixed detector, the positions of a movable piston are found for minimum readings with and without the sample. From the piston displacement and the values of the minima the admittance of the shunt is determined and hence, by means of simple formulas, the dielectric constant of the sample is calculated.

621.317.335.3.029.64 2068

Results of Measurements of Dielectric Constants at 9500 mc by a New Method—S. Le Montagner and J. Le Bot. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 593–594; Feb. 9, 1953.) Using the method previously described (2067 above, Le Bot and Le Montagner) measurements were made on polyethylene, plexiglas and nonpolar and polar liquids, including water. Results are tabulated and compared with figures obtained by other workers. The sensitivity of the method is such that the addition of even 1% of water to pure dioxan is clearly detectable.

621.317.335.3.029.64 2069

The Theory of Measurements of Dielectric Constants at Centimeter Wavelengths—N. V. Kotosonov. (*Zh. Tekh. Fiz.*, vol. 22, pp. 530–536; March 1952.) A general formula (6) is derived for the reflection coefficient of a two-layer dielectric filling a waveguide. From this, assuming that the second layer is air, another formula (7) is derived for the reflection coefficient for any position of the sample in the waveguide. Two methods for measuring the dielectric constant of layers of liquid and solid dielectrics are proposed and optimum conditions of measurement are established.

621.317.336.015.7:621.315.212 2070

The Response to Television and Testing Pulses of Cables with Nonuniform Characteristic Impedance—H. Kaden. (*Arch. elekt. Übertragung*, vol. 7, pp. 157–162, 191–198; March/April 1953.) The effect of the nonuniformities is that the cable output corresponding to a step-voltage input is distorted into a tailing-off form. The tail is due to (*a*) double internal reflections and (*b*) reflections at the cable ends in combination with single internal reflections. Effect (*a*) is proportional to the product of the nonuniformity function and its second derivative; it varies linearly with time for short cables and exponentially for long cables. Effect (*b*) depends on the first derivative of the nonuniformity function; it varies as the square root of the time for short cables and exponentially for long cables. Methods are described for measuring the nonuniformity function and its first and second derivatives; pulses of alternate polarity are used. The correlation range of the nonuniformities is determined; the smaller its value, the longer is the pulse tail. The relation between the autocorrelation function of the nonuniformities and the frequency spectrum of the reflected pulse is discussed in an appendix.

621.317.34.029.63.64 2071

Measurement of the Centimeter-Wave Propagation Coefficient of Quadripoles—W. Klein. (*Fernmeldetech. Z.*, vol. 6, pp. 25–33; Jan. 1953.) Methods of measuring the attenuation and phase coefficients of linear passive quadripoles at centimeter wavelengths are described. Limits are defined within which the node-displacement method of Weissfloch (711 of 1943) is applicable to low-loss quadripoles. Transmission characteristics of a 4-cavity filter for 2.03 kmc, with a 30-mc pass band, are shown. The determination of input impedance from the s.w.r. and the open- and short-circuit impedances is illustrated, with reference to the Smith diagram for the above filter and for a travelling-wave-valve helix.

621.317.341:621.315.212.2 2072

Measurement of the Attenuation of Coaxial Lines (from 1 to 4000 mc)—H. Jassin. (*Câbles & Trans.*, vol. 7, pp. 16–27; Jan. 1953.) Three measurement techniques are described: (*a*) frequencies up to 40 mc, use of a HF bridge; (*b*) 50–400 mc, comparison of output powers of different lengths of identical line (for equal inputs); (*c*) above 500 mc, measurement of standing-wave ratios. Typical attenuation/frequency curves for rigid and for flexible lines are shown.

621.317.35 2073

A Highly Selective Frequency-Spectrum Analyzer—H. Lange. (*NachrTech.*, vol. 2, pp. 471–473, 476; Dec. 1952.) Description and circuit details of an instrument for frequency analysis in the range 30 cps–20 kc. A frequency is selected by adjustment of a 60–80-kc Colpitts oscillator. The difference-frequency signal derived is applied successively to two differential bridge circuits each containing a 60-kc crystal. Amplitudes of 20 mV–100 V can be measured; at 1 V, maximum error is about 1%.

621.317.4.029.5:538.221 2074

High-Frequency Calibration of Magnetic Materials—(*Engineer* (*London*), vol. 195, pp. 329–330; Feb. 27, 1953.) A brief account of National Bureau of Standards methods for determining the RF permeability and loss factor of ferrites and iron powders in the frequency range 50 kc–30 mc. The primary calibration standard is a nonmagnetic coaxial line of variable length and high dimensional accuracy; the magnetic characteristics of a sample are determined from measurements of the lengths of line required to give the same value of inductance with and without the sample inserted. A secondary standard rugged enough to be used in a factory production line is an RF permeameter with a transformer comprising a reference toroid acting as primary and a short coaxial line acting as secondary.

621.317.7:621.396.82 2075

Aircraft Radio-Interference Measurements—M. M. Newman, R. C. Schwantes and J. R. Stahmann. (*Elect. Eng., N. Y.*, vol. 72, pp. 36–40; Jan. 1953.) A double-beam high-speed oscillograph suitable for measurements of either atmospheric or man-made noise is described. One channel has a 10 mc amplifier, the other a 0.1–250 mc distributed-type wide-band amplifier. A 25-kV intensification potential is provided to facilitate photography of the trace. Two methods for more rapid noise analysis are briefly mentioned.

621.317.715 2076

A Moving-Coil Galvanometer of Extreme Sensitivity—K. Copeland, A. C. Downing and A. V. Hill. (*Jour. Sci. Instr.*, vol. 30, pp. 40–44; Feb. 1953.) The galvanometer movement is strongly overdamped, but the short natural period of the coil compensates for this, and full deflection is reached in a few seconds. The very light Cu coil, wound on a Cu frame, is suspended by flat CdCu wires kept taut by two flat-rolled springs, a construction facilitating careful balancing. Hill's photoelectric deflection amplification method (3175 of 1948) is used. Tests show that the mean of 7 readings is reasonably certain to be accurate to within 4×10^{-11} A under normal laboratory conditions.

621.317.715 2077

The Brownian Fluctuations of a Coupled Galvanometer System—A. V. Hill. (*Jour. Sci. Instr.*, vol. 30, pp. 44–45; Feb. 1953.) When two galvanometers are coupled by a photocell and amplifier combination, the fluctuations of the primary instrument are reduced by the inertia and damping of the second. A formula is derived for the rms value of the fluctuations observed on the second galvanometer and is applied to calculations for particular cases.

621.317.725 2078

A High-Resistance Direct-Voltage Valve Voltmeter with a Measurement Range of −500 to +500v—A. Ehmert and R. Mühleisen. (*Z. angew. Phys.*, vol. 5, pp. 43–47; Feb. 1953.) The instrument described has an input impedance of about $5\times10^{12}\Omega$ and an input capacitance of about 0.3 pF; it is suitable for measurements of current down to about 10^{-13} and determinations of capacitance down to 0.1 pF. The calibration curve is practically linear, the slope being 125 μA/500V. The equipment is mains operated, and is compensated against mains voltage fluctuations.

621.317.725:621.396.645 2079

An Electrometer Valve Voltmeter of Wide Range—A. W. Brewer. (*Jour. Sci. Instr.*, vol. 30, pp. 91–92; March 1953.) A dc voltmeter with a range of +500V and a rapid response suitable for atmospheric-electricity measurements is obtained by modifying Farmer's circuit (3337 of 1942) so that the electrometer valve and the associated ordinary valve both act as cathode followers, two further ordinary pentodes being used as "tail valves."

621.317.76.029.63:621.396.615.14 2080

U.H.F. Grid-Dip Meter—A. E. Hylas and W. V. Tyminski. (*Electronics*, vol. 26, pp. 175–177; April 1953.) Oscillators for the frequency range 390–1000 mc are discussed. This rang lies above that of self-oscillation of the valve used, and is attained by means of series-capacitance tuning as described by Pettit (2452 of 1950). The capacitance is distributed to form a balanced circuit, thus providing a maximum-current point on the probe at a location which remains fixed as the oscillator is tuned.

621.317.799:621.396.822.029.64 2081

Noise Comparator for Microwaves—J. J. Freeman. (*Radio & Telev. News, Radio-Electronic Eng. Sec.*, vol. 49, pp. 11, 49; March 1953.) Secondary standards of noise power used at the National Bureau of Standards consist of klystrons or fluorescent tubes; these are calibrated in terms of the noise power of a matched load at a known temperature by means of a "noise bridge" which is a modified form of Dicke's radiometer (475 of 1947). The method of calibration is described, with details of the microwave switch, quarter-wave plate and waveguide transducer used for calibration at 9 kmc.

621.395.625(083.74):621.396.822 2082

Standards on Sound Recording and Reproducing: Methods of Measurement of Noise, 1953—(PROC. I.R.E., vol. 41, pp. 508–512; May 1953.) Standard 53 IRE 19 S1.

621.396.622.6:621.317.3 2083

Testing U.H.F.-TV Mixer Crystals—N. DeWolf. (*Electronics*, vol. 26, pp. 156–160; April 1953.) To evaluate the effect of Ge-crystal mixers on the over-all noise figure of receivers, the conversion loss and noise temperature must be determined. For simplicity, the wide-band conversion loss as defined by Torrey and Whitmer (2989 of 1948) is adopted. Laboratory equipment is described consisting essentially of an UHF receiver adapted for the measurement of conversion loss, noise temperature and RF and IF crystal admittances. Local-oscillator frequency is 900 mc and the IF is 30 mc. A Wollaston wire bolometer is used for the output-measurements. A complete set of measurements can be made in 5 min. Simpler AF methods for production testing are also described.

534.321.9.001.8:532.574 2084

Electronic Flowmeter—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 37, pp. 30–31; Feb. 1953.) The principle of the instrument described is the phase comparison of ultrasonic waves before and after transmission over a fixed distance in a fluid. Transmitter and receiver are identical magnetostriction or piezoelectric transducers, and their connections to the phase meter are interchanged periodically by a rotary commutator or electronic switch. The difference between the two phase shifts is a measure of the flow velocity. Applications include the measurement of air currents and of slow or rapid flow of fluids in pipes.

534.321.9.001.8:538.652:534.232 2085

A High-Frequency Reciprocating Drill—E. A. Neppiras. (*Jour. Sci. Instr.*, vol. 30, pp. 72–73; March 1953.) A high-Q magnetostriction transducer is used in conjunction with a step-up velocity transformer to produce intense vibrations at a low ultrasonic frequency, for drilling or cutting hard brittle materials. See also 1988 of 1952 (Kuris).

538.56.029.63:534.222.2 2086

Reflection of Microwaves by Explosion-Wave Fronts—B. Koch. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 661–663; Feb. 16, 1953.) A sharp discontinuity of the electron density of the medium occurs at the front of an explosion wave. Measurements of the velocity of such a wave have been made by means of the Doppler effect for 23-cm waves reflected from the wave front. The mean value of the velocity thus determined is about 3.5% greater than that given by classical methods. A possible explanation of the discrepancy is given.

621.316.7 2087

Processes in Regulating Systems with Time Delay—K. Küpfmüller. (*Arch. elekt. Übertragung*, vol. 7, pp. 71–78; Feb. 1953.) An investigation is made of the processes consequent on a disturbance of equilibrium in static and astatic systems. Rules for optimum design are derived.

621.38.001.89:621.95 2088

Electronic Prevention of Drill Breakage—(*Engineering* (*London*), vol. 175, pp. 278–279; Feb. 27, 1953.) To prevent breakages when drilling deep holes in Al, a device is used which reverses the drill when the torque exceeds a given amount. The torque-control unit consists essentially of an unbalanced Wheatstone-bridge circuit in which the galvanometer is replaced by the grid circuit of a valve.

621.384.6 2089

Ion Optics in Long High-Voltage Accelerator Tubes—M. M. Elkind. (*Rev. Sci. Instr.*, vol. 24, pp. 129–137; Feb. 1953.)

621.384.611 2090

Phase Stability of the Microtron—C. Henderson, F. F. Heymann and R. E. Jennings. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 41–49; Jan. 1, 1953.) The permissible variation in energy or magnetic field for stable acceleration in an electron cyclotron is calculated and the results are discussed in relation to resonant-cavity design.

21.384.611/.612 2091

Perturbations in the Magnetic Deflector for Synchro-cyclotrons—K. J. Le Couteur. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 25–32; Jan. 1, 1953.)

621.384.612 2092

The Proton Synchrotron at Birmingham University—(*Engineer* (*London*), vol. 195, pp. 271–274, 305–307; Feb. 10 and 27, 1953.) Outline description of design and construction. See also 700 of 1951 (Hibbard).

621.384.612 2093

A 35-Million-Volt Synchrotron—O. Wernholm. (*Ark. Fys.*, vol. 5, Parts 5/6, pp. 565–580; Oct. 26, 1952.) Description of the construction and operation at the Royal Institute of Technology, Sweden, of a synchrotron producing an electron beam with energies continuously variable from 3 to 35 MeV.

621.385.833 2094

Potential and Field of a Cylindrical Electrostatic Lens with Three Slits—M. Laudet. (*Cah. Phys.*, pp. 73–80; Jan. 1953.) Analysis for a system comprising three pairs of half-planes at different potentials.

621.385.833 2095

An Electrostatic Single Lens permitting Rigorous Calculation—P. Schiske. (*Nature* (*London*), vol. 171, pp. 443–444; March 7, 1953.)

621.385.833 2096

Determination of the Trajectory of an Electron by Successive Differentiations—É. Durand. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 236, pp. 471–473; Feb. 2, 1953.)

621.385.833 2097

Astigmatism of Electron Lenses—S. Leisegang. (*Optik, Stuttgart*, vol. 10, pp. 5–14; 1953.)

621.385.833 2098

Investigation of the Asymmetry of [electron] Lenses with Magnetic Pole-Pieces using Rays with Slightly Subtelescopic Paths [im schwach unterteleskopischen Strahlengang]—F. Lenz and M. Hahn. (*Optik, Stuttgart*, vol. 10, pp. 15–27; 1953.)

621.385.833 2099

A Simple Approximate Formula for the Field Distribution along the Axis of Magnetic Electron Lenses with Unsaturated Pole-Pieces—H. Bremmer. (*Optik, Stuttgart*, vol. 10, pp. 1–4; 1953.)

621.385.833 2100

The Adjustment of Spherically Corrected Electron-Optical Systems—R. Seeliger. (*Optik, Stuttgart*, vol. 10, pp. 29–41; 1953.)

621.385.833 2101

Special Construction Features of an Electrostatic [electron] Microscope for the Laboratory—H. Bethge. (*Optik, Stuttgart*, vol. 10, pp. 137–142; 1953.)

621.385.833:538.221 2102

Remanence in Soft-Iron Circuits of Magnetic Electron Lenses—B. v. Borries, F. Lenz and G. Opfer. (*Optik, Stuttgart*, vol. 10, pp. 132–136; 1953.)

621.385.833:621.311.6 2103

High-Frequency High-Voltage Equipment for [electron] Microscope with Intermediate Accelerator—S. Panzer. (*Optik, Stuttgart*, vol. 10, pp. 107–111; 1953.)

621.387.42 2104

Factors influencing the Life of Self-Quenching Counters—S. S. Friedland and H. S. Katzenstein. (*Rev. Sci. Instr.*, vol. 24, pp. 109–112; Feb. 1953.)

621.387.424 2105

Ethylene and Ethyl Alcohol as Quenching Agents in External-Cathode Geiger Counters—R. L. Chasson and M. L. MacKnight. (*Rev. Sci. Instr.*, vol. 24, pp. 212–213; March 1953.)

621.387.424 2106

Construction of Maze-Type Counters [with external cathode], and their Characteristics from 0° to 50° C—R. Favre and C. Haenny. (*Helv. Phys. Acta*, vol. 26, pp. 53–64; Feb. 15, 1953. In French.)

621.387.464 2107

A Method of Increasing the Effective Resolution of Scintillation Counters—K. G. Standing and R. W. Peelle. (*Rev. Sci. Instr.*, vol. 24, pp. 193–195; March 1953.)

PROPAGATION OF WAVES

538.566 2108

Higher-Order Approximations in Ionospheric Wave Propagation—C. O. Hines. (*Jour. Geophys. Res.*, vol. 58, pp. 95–98; March 1953.) Feinstein's results (2608 of 1950), though mathematically correct, are not directly applicable to the physical problem for which they were intended. Similar treatment, but assuming real solutions only, is here developed. Certain results previously obtained, in particular those concerning resonance, are found for certain orders of approximation, but the general validity of the approach is found to be doubtful.

621.391.3 2109

The Fields of an Electric Dipole in a Semi-infinite Conducting Medium—J. R. Wait and L. L. Campbell. (*Jour. Geophys. Res.*, vol. 58, pp. 21–28; March 1953.) Expressions for the electric and magnetic fields at all points inside the conductor are developed in terms of Thomson's functions, the approximations made implying frequencies <100 kc and observation of the fields at distances much less than the free-space wavelength. Curves are shown plotting each field against a parameter involving conductivity, angular frequency and radial distance from the source. An example for for 30-kc propagation in sea water is worked out.

621.396.11 2110

Ionospheric Storm-Warning Services. Assessment of their Value—C. M. Minnis. (*Wireless Eng.*, vol. 30, pp. 103–108; May 1953.) It is suggested that the usefulness of a set of forecasts should be measured in terms of their economic value for a communication system. A figure of merit is derived which expresses the increase in economic value resulting from use of a set of forecasts. This figure includes a parameter representing the effects of ionospheric disturbances and depending also on the type of system operated and on the geographic location of the circuit; its value is unity for a perfect set of forecasts.

621.396.11 2111

Theoretical Curves for Resonance Interaction between Electromagnetic Waves incident Vertically on the Ionosphere—M. Motzo. (*Nuovo Cim.*, vol. 9, pp. 213–219; March 1, 1952.) Bailey's theory of gyro-interaction (9 of 1939) is extended to the case of vertical incidence. The curves obtained in this case have only a single resonance peak.

621.396.11:[530.145.6:537.311.33 2112

Some Particular Features of Tropospheric Propagation and their Analogies in Wave Mechanics—M. Ortusi. (*Ann. Radioélect.*, vol. 8, pp. 81–95; Jan. 1953.) Classical theory of tropospheric propagation is summarized, and the limits within which it is valid are discussed. The problem is shown to be mathematically analogous to that of determining particles trajectories in wave mechanics. The waves associated with the electron and hole aspects of the elementary particle are shown to correspond with particular cases of propagation in Gamow and Eckersley modes. The analogy is developed by comparing the form of a potential barrier separating solid media with the refractive-index profile for an atmospheric column including an inversion layer. The discussion indicates the manner in which holes are created inside barrier layers.

550.372.621.396.11 2113

Effect of a Large Dielectric Constant on Ground-Wave Propagation—J. R. Wait and L. L. Campbell. (*Canad. Jour. Phys.*, vol. 31, pp. 456–457; March 1953.) Curves are plotted showing variation of field strength with distance from transmitter for values of ground dielectric constant ranging from 2 to 200, assuming a value of 0.1×10^{-4} emu for the ground conductivity. Vertically polarized radiation of frequency 200 kc is assumed. At great distances from the transmitter, high values of dielectric constant give rise to great increase of field strength.

621.396.11:551.510.535 2114

Magneto-ionic Multiple Splitting determined with the Method of Phase Integration—W. Pfister. (*Jour. Geophys. Res.*, vol. 58, pp. 29–40; March 1953.) The method used involves a first-order W.K.B. approximation with an integration path in the complex plane. The propagation constant is represented on a four-sheeted Riemann surface corresponding to up-going and down-coming waves of ordinary and extraordinary types of polarization. The branch points connecting the sheets represent the reflection points or the coupling points. When suitable integration paths around the branch points are chosen, five fundamental magneto-ionic components are found. Numerical calculations for moderate magnetic latitude based on an ionosphere model with Chapman distribution of electrons and exponential decrease of collision frequency are made and the results presented in curves showing virtual height and absorption for the five components found.

621.396.11:551.510.535 2115

Ionosphere Critical Frequencies at Oblique Incidence—H. Poeverlein. (*Z. angew. Phys.*, vol. 5, pp. 15–19; Jan. 1953.) The dependence of muf on the inclination of the geomagnetic field and on the direction of wave propagation is investigated on the basis of previously prepared graphs showing (*a*) the ratio of critical frequency at vertical incidence to that at oblique incidence, as a function of angle of incidence, and (*b*) the ratio of critical frequency without magnetic field (identical with that of the ordinary ray in the E-W plane) to that of the ordinary ray in the plane of the magnetic meridian, also as a function of the angle of incidence. For low frequencies (0.75 mc) the latter ratio has a rather high value at the magnetic equator.

621.396.11.029.55 2116

Observations at Calcutta of Pulses Transmitted from Delhi—S. S. Baral and A. K. Saha. (*Indian Jour. Phys.*, vol. 26, pp. 521–538; Nov. 1952.) An account of observations during November 1950 and May–June 1951 of transmissions at frequencies of 17.74 and 21.7 mc. Possible modes of propagation are discussed; it is concluded that the signals are received by single reflection from the F_2 layer and that multiple reflections from the F_2 layer or single reflections from the E and F_1 layers are not involved. The signal attenuation in the Delhi-Calcutta path is estimated and the results are tabulated. Statistical analysis of the fluctuations of the received signals indciates greater stability for the lower ordinary ray than for the upper ordinary (Pedersen) ray.

621.396.11.029.55 2117

Extended-Range Radio Transmission by Oblique Reflection from Meteoric Ionization—O. G. Villard, Jr., A. M. Peterson, L. A. Manning and V. R. Eshleman. (*Jour. Geophys. Res.*, vol. 58, pp. 83–93; March 1953.) Tests show that communication can be maintained between low-power stations approximately 1200 km apart at frequencies around 14 mc when no detectable F-, E- or sporadic-E-layer reflections are taking place. The average signal level is 10 db above noise level, and the signal appears to come from the direction of the transmitter, a continual small shift suggesting that there is no single point of reflection. Tests involving three transmission paths of differing lengths were made to determine total duration of discernible meteor-trail echoes. Results showed increased echo durations over oblique paths and thus support the theory that the signals are obliquely reflected from diffuse meteor trails.

621.396.11.029.62/.63 2118

Graphs relating to the Propagation of Ultrashort Waves between Points within Visible Range—L. Sacco. (*Poste e Telecomunicazioni*, vol. 21, pp. 55–71; Feb. 1953.) Various graphs are presented for facilitating design calculations for usw and microwave links, and for determining the best wavelength from the point of view of efficiency and stability of operation in relation to meteorological and geographical factors.

621.396.11.029.62:551.510.535 2119

Scattering of 56-mc Radio Waves from the Lower Ionosphere—P. A. Forsyth, B. W. Currie and F. E. Vawter. (*Nature* (*London*), vol. 171, pp. 352–353; Feb. 21, 1953.) Radar records made during 1952 at Saskatoon show persistent scatter echoes with a diurnal and seasonal variation. Similar observations at Ottawa by McKinley and Millman on 33 mc are also mentioned (see 2000 above.)

621.396.812.3 2120

Some Studies on Random Fading Characteristics—R. B. Banerji. (*Proc. Phys. Soc.* (*London*), vol. 66, pp. 105–114; Feb. 1, 1953.) Critical examination of the theory of random signals reflected from the ionosphere indicates that the velocity distribution of fading is independent of the power spectrum of the reflected wave. The effect produced on the velocity distribution of fading by superposing a steady signal on a randomly fading wave is found to be small, resulting in only a very slight increase of the velocity of fading. An alternative method is given for determining the autocorrelation function of the fading pattern, from which the power spectrum of the reflected wave can be derived by inverse Fourier transformation. This provides an alternative method of determining the velocity of drift of ionospheric irregularities.

RECEPTION

621.396.62 2121

The Design of a Comparator A.M./F.M. Broadcast Receiver—F. H. Beaumont. (*Jour. Brit. Inst. Radio Engs.*, vol. 13, pp. 131–148; March 1953.) Discussion of the design of a superheterodyne receiver to meet a specification drawn up by the BBC for making subjective comparisons of AM and FM transmissions in the frequency band 87.5–95 mc. The choice of circuits for RF amplifier, frequency changes, IF amplifier, detector, discriminator, noise limiters and AF stages is dealt with in detail.

621.396.62:621.396.664 2122

An Automatic-Scanning Receiver for Radio Monitoring—F. J. M. Laver, F. M. Billinghurst and F. J. Lee. (*P. O. Elect. Eng. Jour.*, vol. 45, Part 4, pp. 149–153; Jan. 1953.) The tuning frequency of a sensitive communication receiver is swept over a predetermined band of width up to 1 mc once every two minutes by means of a drive unit mechanically coupled to the tuning spindle. A continuous permanent record is obtained on electrosensitive paper of all signals received above a given strength in any desired frequency band between 4 and 27.5 mc.

621.396.62:621.396.812 2123

Highly Stable Receiver for Space Waves—G. Bartels. (*NachrTech.*, vol. 3, pp. 28–29; Jan. 1953.) For investigations of the lower ionosphere, field-strength measurements are made on long-wave transmissions over paths of some hundreds of kilometers. A simple 4-valve receiver is used comprising HF stage, mixer with quartz-controlled oscillator, IF amplifier and crystal detector, and dc valve voltmeter bridge with recording instrument. The sensitivity is about 0.3 mV/m for full-scale deflection of a 1 mA/150 mV instrument.

621.396.621:621.396.822:519.272 2124

The Detection of Weak Signals by Correlation Methods—P. Rudnick. (*Jour. Appl. Phys.*, vol. 24, pp. 128–131; Feb. 1953.) Comparison of an autocorrelation system with a conventional detection system comprising band-pass filter, rectifier and low-pass filter indicates that for the same detection threshold the former system has to be far more complex. An alternative system involving multiplication of the incoming signal by a local sinusoidal signal is shown to be closely parallel to the conventional system.

621.396.621.029.55:621.396.822 2125

Dependence of the Signal/Noise Ratio of Short-Wave Receivers on the Input Voltage—K. Fischer. (*Telefunken Ztg.*, vol. 26, pp. 43–48; Jan. 1953.) In receivers with agc it is preferable, so far as signal/noise ratio is concerned, for the gain control to be effected in the later stages, but from the point of view of cross modulation the regulation should be effected at the receiver input. In order to determine the best compromise between these conflicting requirements, experiments were carried out with circuit arrangements in which the gain control could be applied to different groups of valves. With such arrangements cross-modulation can be reduced with an accompanying reduction of the signal/noise ratio, or the latter increased at the expense of cross-modulation sensitivity, so that the optimum circuit for particular conditions can be adopted.

621.396.621.54 2126
Problems concerning Tracking Calculations in the Superheterodyne—J. Mohrmann. (*Funk u. Ton*, vol. 7, pp. 1–9; Jan. 1953.) Critical discussion of tracking methods, underlining errors and omissions involved. See also 2323 of 1952.

621.396.621.54:621.396.822 2127
Noise Factor and its Relation to the Sensitivity of a Superheterodyne Receiver with Crystal Mixer—E. Willwacher. (*Telefunken Ztg*, vol. 26, pp. 33–42; Jan. 1953.) Noise factor is defined and a formula is derived for the noise factor of a series arrangement of *n* quadripoles, each of which may add to the input noise. Analysis is presented for a mixer stage, an equivalent circuit being utilized. The results enable the optimum sensitivity of a superheterodyne receiver to be calculated. Numerical calculations are in good agreement with experimental values.

621.396.622.6:621.317.3 2128
Testing U.H.F.-TV Mixer Crystals—DeWolf. (See 2083.)

621.396.622.71 2129
Calculation of the Efficiency and Damping of Diode Detectors—H. H. van Abbe. (*Electronic Applic. Bull.*, vol. 13, pp. 65–71; May 1952.) See 1195 of April.

621.396.662.029.63 2130
Automatic Tuning Devices in Microwave Receivers—G. Voigt. (*NachrTech.*, vol. 2, pp. 456–459; Dec. 1952.) The principles of an automatic system are described whereby the mean IF is maintained at the center frequency of the demodulation characteristic. The control voltage is derived from the dc component of the discriminator. Methods of correcting for frequency fluctuations (*a*) up to ±250 kc, (*b*) up to 0.5–1 mc, are outlined.

621.396.828 2131
Eliminating Interference caused by Railway Systems fed with Single-Phase Current at 15 kV—J. Meyer de Stadelhofen. (*Tech. Mitt. schweiz. TelephVerw.*, vol. 31, pp. 33–52; Feb. 1, 1953. In French.) Report of investigations made by the Swiss Post Office and Swiss Railways to determine the extent of radio interference caused by the electric traction system and to find means for reducing it. As regards the Sottens, Beromünster and Monte Ceneri stations, operating in the frequency range 500–800 kc, the interference can be reduced to 1/10–1/30 of its original value by simple means.

621.396.828 2132
A "Codan" for A.M. Receivers—J. B. Rudd. (*Proc. I.R.E.* (*Australia*), vol. 14, pp. 33–40; Feb. 1953.) The "Codan" (carrier-operated-devices-anti-noise) is an auxiliary circuit device which selects a wanted carrier out of noise at an IF stage by means of a narrow-band filter and operates a relay to open the audio circuit. It can operate with signal/noise ratios >1. A practical unit is described which uses a double-heterodyne process and a band-pass delay line to overcome the effects of frequency drift. Performance figures are given.

621.396.828:621.327.43 2133
Suppression of Radio Interference from Fluorescent Lamps over the Frequency Range 150 kc–1.5 mc—J. Meyer de Stadelhofen. (*Tech Mitt. schweiz. Telegr.-TelephVerw.*, vol. 30, pp. 239–248; Aug. 1, 1952. In French and German.) RF processes in fluorescent-lamp circuits are analyzed. For satisfactory suppression of interference, both a mains filter and a screen round the tube are essential. The ballast coil should be arranged symmetrically and should have low-capacitance windings. Measurement results indicate that interference can be reduced to a tenth of its original value, even in unfavorable circumstances, using simple means.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 2134
Symposium on Information Theory, London, September 1950—(*Trans. I.R.E.*, PGIT-1, 208 pp.; Feb. 1953.) Full report of the proceedings at the symposium noted in 984 of 1951 (Jackson).

621.39.001.11 2135
The Information Content of an Electromagnetic Signal—D. Gabor. (*Arch. elekt. Übertragung*, vol. 7, pp. 95–99; Feb. 1953.) The amount of information that can be extracted from the signal by means of a valve is limited by (*a*) spontaneous fluctuations of the signal, (*b*) shot effect, and (*c*) a phenomenon resulting from the quantum nature of the em field, in accordance with which an electron interchanging energy with the signal field affects the phase of the field by an indeterminate amount, even when the resultant energy exchange is zero. The amount of information obtainable by simultaneous measurement of amplitude and phase is thus of an order no greater than would be obtained by means of an ideal photon counter, which is insensitive to phase changes. See also 208 of 1951.

621.39.001.11 2136
Optimum Linear Shaping and Filtering Networks—R. S. Berkowitz. (Proc. I.R.E., vol. 41, pp. 532–537; May 1953.) Analysis is made to derive the optimum transfer characteristics for the transmitter and receiver of a linear communication system, assuming power levels and over-all distortion to be fixed. Formulas are derived based on two different criteria, namely (*a*) the mean-square value of the receiver-output noise component, and (*b*) the probability that the receiver-output-noise component exceeds a given value at least once during a given observation period. The choice of criterion appropriate in a particular case depends on the mechanism by which the receiver output is converted into information.

621.392.001.11 2137
Analytical Signals with Limited Spectrum: Part 2—J. A. Ville. (*Câbles & Trans.*, vol. 7, pp. 44–53; Jan. 1953.) Applications of the theory developed previously (1020 of 1950) are made to the determination of the relation between phase change and attenuation in a linear network as a function of frequency and to analysis of the limiting distortion compensation practicable in the presence of noise. The synthesis of a signal by superposition of $\cos^2$-type signals (2933 of 1951) is related to the problem of resolution of a signal by means of Shannon's formula; the synthesis can be simplified by referring signals of the $\sin t/t$ type to $\cos^2$-type signals by suitable regrouping of terms. See also 1177 of April (Jouguet).

621.394.324 2138
Progress in Teleprinter Technique—K. Reche. (*Elektrotech. Z., Edn. A*, vol. 74, pp. 4–10; Jan. 1, 1953.) Recent developments resulting in faster, more certain and more economical operation are described, with details of typical equipment.

621.396.029.6:621.396.8:519.2 2139
Problems of U.S.W. Coverage—F. von Rautenfeld and H. W. Fastert. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 9–17; Jan./Feb. 1953.) Methods previously considered for determining the effective service area of a vhf transmitter [213—Gressmann and Kaltbeitzer, and 214—Grosskopf, of January] are illustrated by detailed calculations for the fm transmitters at Lingen and Nordhelle, 156 km apart, for both co-channel and adjacent-channel operation. See also 197 of January (Bangen and Fastert).

621.396.43+[621.396.65:621.397.26 2140
The State of Development of Directional Radio Equipment and the Planning of Directional Radio Links in the German Federal Republic—K. O. Schmidt. (*Fernmeldetech. Z.*, vol. 6, pp. 51–58; Feb. 1953.) Illustrated review of features of television relay equipment and multichannel fm and ppm telephony systems.

621.396.619.13:621.396.41 2141
Calculation of the Distortion of a Frequency-Modulated Wave—J. P. Vasseur. (*Ann. Radioélect.*, vol. 8, pp. 20–35; Jan. 1953.) Formulas are derived which are applicable for calculating both the distortion due to variation with frequency of the delay produced by an amplifier, and that due to the superposition on the main signal of echo signals caused by reflection in aerial feeders, etc. The analysis applies to both low and high modulation rates and to both single-frequency and composite signals. In multiplex systems frequency distortion may give rise to crosstalk; a formula due to Lewin (986 of 1951) is used to calculate the magnitude of this effect for the case of slow modulation.

621.396.619.16 2142
Pulse Distributors—H. Oberbeck. (*Telefunken Ztg*, vol. 26, pp. 23–32; Jan. 1953.) In multichannel PM systems, different types of equipment for distributing the pulses to the various channels are required according to the type of modulator and demodulator used. Descriptions are given of the basic principles and special features of various types of pulse distributor, detailed accounts of which have been published previously.

621.396.65 2143
Radio-Telephone Link between Turkey and the West—(*Engineer* (*London*), vol. 195, p. 363; March 6, 1953.) Brief details of transmitter and receiver installations at Ankara providing a new radiotelephone link with Western Europe and the U.S.A., and an extension of existing radiotelegraph facilities.

621.396.65:621.396.41:621.396.619.24 2144
Single-Sideband Multichannel Operation of Short-Wave Point-to-Point Radio Links: Part 1—General Survey—W. J. Bray and D. W. Morris. (*P. O. Elect. Eng. Jour.*, vol. 45, part 3, pp. 97–103; Oct. 1952.) The principles of ssb working are outlined, its advantages are pointed out and the basic techniques used for multichannel telegraphy and telephony are described.

621.396.65:621.396.61/.62 2145
Single-Sideband Multichannel Operation of Short-Wave Point-to-Point Radio Links: Part 2—Owen and Ewen. (See 2175.)

621.396.65:621.396.619.16 2146
Transmission Performance Figures for Directional PPM Systems—P. Barkow. (*Fernmeldetech. Z.*, vol. 6, pp. 2–11; Jan. 1953.) Tests have been conducted on PPM systems operating in Germany. Performance figures are compared with CCIF specifications based on data for carrier-current systems. Noise figures of equipment for long-distance and local traffic are analyzed and the permissible noise levels throughout a communication chain are discussed in relation to transmitter power, receiver sensitivity and bandwidth.

621.396.65:621.397.6 2147
The Decimeter-Wave Beam Radio Equipment FREDA I—Behling, Brühl and Willwacher. (See 2160.)

621.396.712+[621.396.712:621.396.66 2148
Transmitting and Monitoring Stations of the NWDR—(*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 18–21; Jan./Feb. 1953.) A list is given of the NWDR medium-wave, sw, usw and television transmitting stations, as at 1st January 1953, with details of height above sea level, frequency, type of modulation, power, type of aerial, programme, and date of commencing operation. The equipment of the monitoring stations at Wittsmoor, Hamburg and Norderney is also listed.

621.396.72.029.62 2149
Transmitter and Studio for Private USW Stations—Tetzner. (See 2178.)

SUBSIDIARY APPARATUS

651-526 2150
Synthesis of Servomechanisms by Root Locations—D. W. Russell and C. H. Weaver. (*Elec. Eng., N. Y.*, vol. 72, p. 41; Jan. 1953.) Summary only.

621.311.6 2151
The Cockcroft-Walton Voltage-Multiplying Circuit—E. Everhart and P. Lorrain. (*Rev. Sci. Instr.*, vol. 24, pp. 221–226; March 1953.) The original circuit (see *Proc. Roy. Soc. A*, vol. 136, p. 619; 1932) is studied by considering it as a transmission line. Losses are examined, and formulas are developed for the voltage efficiency for a given size of capacitor and number of stages. Two modifications to improve voltage efficiency are discussed; the first is the use of a loading coil at the hv end of the line, the other is the inclusion of inductors in series with each of the capacitors.

621.311.62 2152
Improved Variable Power Supply—W. Creviston. (*Radio & Telev. News*, vol. 49, pp. 42, 115; April 1953.) A modification of Walker's circuit (3562 of 1952) is described which provides direct-voltage outputs from 50 to 300 V, together with heater supplies. Inherent regulation is as good at 50 V as at 300 V.

621.314.632 2153
Hermetically Sealed Magnesium Copper-Sulfide Rectifiers—M. Gamble. (*Elec. Mfg*, vol. 48, p. 132; Oct. 1951.) Short note on the construction methods used for this type of rectifier, which has an operating range from −70° to +200° C or higher.

621.314.632/.634 2154
Advanced Developments in Metallic Rectifiers—W. F. Bonner and F. J. Oliver. (*Elec. Mfg*, vol. 48, pp. 128–133, 288; Oct. 1951.) Discussion of improvements relative to higher operating voltage, smaller size, operation at high and at low temperatures, and higher efficiency, largely resulting from requirements of the Services. Se rectifiers are particularly considered.

621.314.671 2155
New Rectifier Tube for Extremely High Power and Voltage Levels—T. H. Rogers. (*Elec. Eng., N. Y.*, vol. 72, pp. 51–56; Jan. 1953.) This tube has a Th-W filament with catenary configuration and a cylindrical anode of Ta coated with W powder. Its characteristics are: anode dissipation rating, 1.5 kW; peak current rating, 10_A; maximum peak inverse voltage, 110 kV.

621.315.616:621.355.2 2156
Microporous Thermoplastic Separators for Lead Acid Batteries—(*Engineer (London)*, vol. 195, pp. 386–387; March 13, 1953.) Brief account of the manufacture and properties of "Porvic," a PVC sheet material. A similar account is given in *Plastics*, vol. 18, pp. 104–106; April 1953.

621.355.2 2157
Addition Agents for Negative Plates of Lead-Acid Storage Batteries: Part 2—Pure Organic Compounds—E. J. Ritchie. (*Jour. Electrochem. Soc.*, vol. 100, pp. 53–59; Feb. 1953.) The best results were obtained with carbohydrates and some of the homologous phenolic compounds.

TELEVISION AND PHOTOTELEGRAPHY

621.317.336.015.7:621.315.212 2158
The Response to Television and Testing Pulses of Cables with Nonuniform Characteristic Impedance—Kaden. (See 2070.)

621.397.26:621.396.65]†621.396.43 2159
The State of Development of Directional Radio Equipment and the Planning of Directional Radio Links in the German Federal Republic—Schmidt. (See 2140.)

621.396.65 2160
The Decimetre-Wave Beam Radio Equipmen FREDA I—H. Behling, G. Brühl and E. Willwacher. (*Telefunken Ztg*, vol. 26, pp. 4–22; Jan. 1953.) The name FREDA signifies "frequenzmodulierte Dezimeteranlage." The equipment is designed for the Hamburg-Cologne link, the operating frequency being 1755 +*n*.60 mc for the different channels. A description is given, with block diagrams, of the arrangements at the terminal transmitting and receiving stations and at a relay station. Details of the construction of the transmitter HF oscillator, mixer and amplifier stages are illustrated by photographs. Modulator and demodulator schematic circuit diagrams are given and performance data are summarized. Transmitter power is 5 W and the gain of the 3-m parabolic reflector, with dipole feed, relative to an elementary dipole, is 1060. Signal/noise ratio for the television channel is good.

621.397.335 2161
Synchronization and Pulse Technique in Television—J. Günther. Correction slip inserted in Nov./Dec. issue of *Tech. Hausmitt. NordwDtsch. Rdfunks* gives a corrected diagram for Fig. 1 of paper abstracted in 1153 of April.

621.397.5:535.623 2162
Recent Advances in Color Television—F. W. de Vrijer. (*Tijdschr. ned. Radiogenoot.*, vol. 18, pp. 105–112; March 1953. In English.) Systems and apparatus developed in the USA are discussed.

621.397.61/.62 2163
Storage and Picture-Difference Methods in Television Reception—F. Schröter. (*Arch. elekt. Übertragung*, vol. 7, pp. 63–70; Feb. 1953.) Use of a storage method at the receiver eliminates the need to transmit "redundant" information, i.e. signals corresponding to picture points which have undergone no change since the previous scan. A description is given of a particular type of storage tube using a 2–3-μ thick insulating target with a mesh-type anode on the scanned face and a mesh-type photocathode on the other face. Such a tube can provide a flicker-free picture of good brightness with a frame frequency of 16 per second. For use in converting pictures brought by line to a broadcasting station, the tube can be modified by including the photocathode in a Farnsworth-type dissector arrangement. Reception by this method combines satisfactorily with transmission by picture-difference methods using two-speed scanning, leading to a substantial reduction of the required bandwidth and transmitter power.

621.397.61 2164
Television Camera Equipment of Advanced Design—L. L. Pourciau. (*Jour. Soc. Mot. Pict. Telev. Engs.*, vol. 60, pp. 166–180; Feb. 1953.) Descriptions are given of the servo circuits and mechanical design features of a television camera chain in which lens selection and focus are remotely controlled. Special emphasis is laid on convenient grouping of controls.

621.397.61:621.396.619.23 2165
Spiral-Beam Tube Modulates 1 kW at UHF—Cuccia. (See 2188.)

621.397.61(494) 2166
Justification of the Choice of the Dôle as Site for a Television Transmitter—H. Laett and J. Dufour. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, vol. 30, pp. 264–270; Sept. 1, 1952.) (In French.) See 246 of January (Laett).

621.397.611.2 2167
Some New Aspects of the Construction and Application of the Image Iconoscope—H. Bruining. (*Le Vide*, vol. 7, pp. 1248–1255; Nov. 1952.) Recent improvements in design are described, including the use of (*a*) a close-mesh grid in front of the photocathode to eliminate ion spots, (*b*) an L-type cathode in the electron gun. A focusing system providing variable magnification is discussed. Methods of target stabilization are noted.

621.397.62 2168
Application of the ECL80 and EQ80 Tubes in Flywheel Synchronization Circuits—A. Boekhorst, P. D. van der Knapp and P. A. Neeteson. (*Electronic Appl. Bull.*, vol. 13, pp. 72–80; May 1952.) Full details are given of two practical arrangements. The first uses a Type-ECL80 triode-pentode in the phase discriminator circuit and effects comparison between pulses at incoming synchronizing frequency and local line-timebase frequency; the second uses a Type-EQ80 enneode and effects comparison between pulses at incoming synchronizing frequency and a sawtooth voltage at local line frequency. Additional features for suppressing the effects of interference on the synchronization are described.

621.397.62 2169
Television Converter—C. A. Marshall. (*Wireless World*, vol. 59, pp. 223–226; May 1953.) Detailed description of a frequency-converter unit for use with nontunable television receivers. The particular values of circuit components given are for adapting a Channel-1 receiver (45mc vision, 41.5 mc sound) to Channel-3 reception (56.75 mc vision, 53.25 mc sound), but the general design is not restricted to these channels.

621.397.62:535.88 2170
Special Problems in Television Large-Picture Installations—E. Schwartz. Correction slip inserted in the Nov./Dec. issue of *Tech. Hausmitt. NordwDtsch. Rfdunks* gives a corrected diagram for Fig. 1 of paper abstracted in 1168 of April.

621.397.621.2:621.385.832 2171
Transfer Characteristics and Mu Factor of Picture Tubes—K. Schlesinger. (Proc. I.R.E., vol. 41, pp. 528–532; May 1953.) The characteristics of television-receiver-tube guns are analyzed, taking account of the variation with grid voltage of the active cathode area. The value of gamma lies between 2.4 and 2.5 for various structures, while the value of the amplification factor is strongly influenced by changes in system parameters. The theorectial results were verified experimentally.

621.397.645.029.62 2172
Fundamental Problems of H.F. and I.F. Amplifiers for TV Reception: Part 3—Feedback and Practical Considerations following on the Theory—Uitjens. (See 1954.)

621.397.82 2173
Television Reception and Interference—W. Werner. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 1–8; Jan./Feb. 1953.) Methods are suggested for reducing interference from cw signals which produce effects in the IF band, and also from impulse voltages. Radiation from the deflection system of the receiver cr tube and from the receiver local oscillator is discussed, practical methods of measuring it are noted and means found effective in reducing it are described. Screening of the picture tube to reduce X-ray emission is also mentioned.

621.397.82 2174
A Combining Unit for Superimposing Two Television Pictures on the Same Cathode-Ray Tube—D. Wray. (*P. O. Elec. Eng. Jour.*, vol. 45, Part 4, pp. 172–174; Jan. 1953.) Apparatus is described by means of which another waveform can be combined with the television signal without loss of synchronism. The effect on the picture of various types of interference cna be simulated. The delay occurring in a long-distance looped television transmission system is evidenced by the displacement between the two pictures observed simultaneously when the signal generated at the transmitter is combined

with that received after travelling round the loop.

TRANSMISSION

621.396.61/.62 2175

Single-Sideband Multi-Channel Operation of Short-Wave Point-to-Point Radio Links: Part 2—F. C. Owen and A. B. Ewen. (*P. O. Elect. Eng. Jour.*, vol. 45, Part 4, pp. 154–159; Jan. 1953.) The present equipment is an improved form of that previously described [2395 of 1948—Bray *et al.*]; it generates a low-power independent-sideband signal comprising two 6-kc channels, one on each side of a reduced-level 3.1-mc pilot carrier, suitable for application to the final modulator and power-amplifier stages of a sw transmitter. Alternatively a single-channel dsb signal can be generated. The associated monitor receiver is designed to accept signals from the transmitter drive unit at 3.1 mc or from the power-amplifier stages at radiation frequency. Part 1: 2144 above (Bray and Morris).

621.396.61.026.446.029.53:621.396.97 2176

Megawatt A.M. Broadcast Transmitter—J. O. Weldon. (*Tele-Tech*, vol. 12, pp. 50–51; 116; Jan. 1953.) Illustrated description of equipment comprising two identical 500-kW transmitters whose outputs are combined to feed 1 MW to the aerial system. Four Machlett 250-kW Type-ML5682 tubes are used in the power amplifier of each transmitter, with four more in the output stage. Grid-bias modulation of the first Type-ML5682 tube is effected by means of four Type-805 tubes in parallel constituting the output stage of the AF amplifier. Operating characteristics in the frequency range 540–1600 kc were found consistently good.

621.396.619.11 2177

The Impulse Modulator—H. Moss. (*Jour. Brit. I.R.E.*, vol. 13, pp. 150–159; March 1953.) A highly linear amplitude modulator is achieved by using a negative-feedback circuit in which the tube anode current is arranged to be as nearly as possible proportional to the anode voltage and applying rectangular on-off switching pulses to the tube grid at carrier frequency, the modulating voltage being applied to the anode. The modulated RF output is taken from a resonant load in series with the tube. 100% depth of modulation can be obtained without requiring residual carrier neutralization. Harmonic distortion, modulation efficiency and linearity, and the influence of pulse mark/space ratio, are discussed.

621.396.72.029.62 2178

Transmitter and Studio for Private USW Stations—K. Tetzner. (*Funk-Technik (Berlin)*, vol. 8, pp. 132–133; March 1953.) Brief description of commercially produced equipment for use in local broadcasting stations owned by cultural institutions etc. in Germany. Transmitter powers range from 50 W to 250 W, the operating frequency being near 100 mc. Studio equipment to satisfy basic requirements is outlined.

VALVES AND THERMIONICS

546.289:621.314.632 2179

Evaporated Point-Contact Rectifiers—E. G. Roka, C. H. Jackson and R. P. Ulrich. (*Jour. Appl. Phys.*, vol. 24, pp. 228–229; Feb. 1953.) A study has been made of the effect on a Ge rectifier of replacing the "whisker" contact by a contact comprising a very small area of evaporated silver. A significant improvement of the rectifying properties is observed. If the rectifier with whisker contact is subjected to alternating pulses of amplitude about 30V, the reverse resistance is thereby increased and the substitution of the evaporated contact then produces no improvement.

546.289:621.385.2 2180

Negative Resistance in Germanium Diodes—J. Kauke. (*Radio & Telev. News, Radio-Electronic Eng. Sec.*, vol. 49, pp. 8–10; April 1953.) Ge diodes show negative-resistance effects at values of the reverse voltage somewhat higher than the rated continuous-working voltage. Applications of such effects in the production of a swatooth-wave oscillator, a sine-wave oscillator, a lock-in circuit and a voltage divider are described.

621.383.27:621.387.464 2181

Photomultipliers Particularly Suitable for counting Scintillations—É. Morilleau, H. Dormont and R. Champeix. (*Compt. Rend. Acad. Sci. (Paris)*, vol. 236, pp. 474–476; Feb. 2, 1953.) Various measures adopted to obtain improved performance are indicated. Curved reflector-type Cu dynodes coated with Cs are used. By focusing the emission from the cathode on to a very small first dynode a low value of background noise is achieved together with constant gain and optical screening of the cathode.

621.385:621.396.615.142 2182

Influence of the Lorentz Force on Space-Charge Waves in Electron Beams—J. Labus. (*Arch. elekt. Übertragung*, vol. 7. pp. 88–94; Feb. 1953.) For a beam of infinite diameter there are two possible values of phase velocity of the space-charge waves. When the formula is modified to take account of finite beam diameter, an infinite number of values of phase velocity are found. On taking into account the effect of the Lorentz force on the motion of the electrons, the formula obtained is very little different from that for the infinite-diameter beam, and gives values of phase velocity nearly independent of beam diameter and nearly equal.

621.385.029.6 2183

Travelling-Wave Tubes—R. Kompfner. (*Rep. Progr. Phys.*, vol. 15, pp. 275–327; 1952.) The historical development of traveling-wave tubes is described and a simple theory of their operation is outlined. More complete theory due to Pierce is sketched which permits calculation of the effects of attenuation, of non-synchronous velocities, and of space charge. Recent developments in the helix-type traveling-wave tube are described, such as Brillouin focusing of electron streams in axial magnetic field and noise characteristics of electron streams. Various types of traveling-wave tube are described and their special features are discussed. Wherever possible, attempts have been made to explain the modes of operation of the various types in simple physical terms. Over 140 references.

621.385.032.216:537.533.8 2184

Changes in Secondary and Thermionic Emission from Barium Oxide during Electron Bombardment—J. Woods and D. A. Wright. (*Brit. Jour. Appl. Phys.*, vol. 4, pp. 56–61; Feb. 1953.) Further experiments on BaO films (see 895 of March) are reported. Changes in secondary emission after 15–30 min bombardment depend on the film thickness and are due to a reduction process forming an excess of metal within the oxide. Thermionic emission also varies with thickness and there is a close correlation between the two types of emission in respect of decay and recovery effects.

621.385.1:621.396.822 2185

Noise in Gas Discharges—A. van der Ziel. (*Jour. Appl. Phys.*, vol. 24, pp. 223–224; Feb. 1953.) Comment on 2670 of 1950 (Parzen and Goldstein) indicating that to make the noise-power formula there derived valid for all frequencies the term corresponding to shot noise requires modification. The theory of shot noise in semiconductors (2040 above) is applied. The revised formula indicates that the shot noise is much greater at low than at high frequencies.

621.396.615.141.2 2186

A 3 cm-Magnetron for Beacons—G. A. Espersen and B. Arfin. (*Philips tech. Rev.*, vol. 14, pp. 87–94; Sept./Oct. 1952.) (See 3184 of 1951.)

621.396.615.141.2 2187

New Magnetron Oscillator with Interdigital Circuit—A. Leblond, O. Doehler and R. Warnecke. (*C. R. Acad. Sci. (Paris)*, vol. 236, pp. 55–57; Jan. 5, 1953.) The oscillatory system of a multislot magnetron can be considered as a section of a line with periodic structure closed on itself [see 1543 of May—Leblond *et al.*]; with such a system, in order to avoid the sudden frequency jump on passing from one oscillation mode to another, the phase velocity of the fundamental wave of the delay line constituted by the oscillatory system should be of opposite sign to the group velocity, and the dispersion should be relatively low. A suitable interdigital structure satisfying these requirements is described with which a peak HF output of 300 kW on 11-cm wavelength has been obtained from a tube operated with peak anode voltage of 22 kV, peak anode current 26A and magnetic field 2,700 gauss.

621.396.619.23:621.397.61 2188

Spiral-Beam Tube Modulates 1 kW at UHF—C. L. Cuccia. (*Electronics*, vol. 26, pp. 130–134; April 1953.) A modified form of the electron-beam coupler tube previously described (2975 of 1949) is used as a television modulator in the 800 mc frequency range. Modulation is performed by means of five auxiliary electron beams in the output cavity; a 50-V modulator grid swing can control power output over a range of 98% of maximum (300W with a beam volatge of 750V). The device has good linearity, a bandwidth of 5 mc and a transfer efficiency of 50%.

621.396.622.63 2189

Conductivity of and Flicker Effect in Crystal Detectors—N. Nifontoff. (*Onde élect.*, vol. 33, pp. 58–61; Jan. 1953.) See 1037 of April and back references.

621.396.622.63:546.28 2190

Silicon Crystal Detectors—J. M. Mercier and R. P. Musson-Genon. (*Onde élect.*, vol. 33, pp. 40–57; Jan. 1953.) An account of the methods of manufacture of Si detectors adopted by the Compagnie Française Thomson-Houston and of their characteristics and use as mixers in uhf circuits. (See also, Mercier, 2070 of 1951.)

MISCELLANEOUS

025.45:621.3 2191

Work of the 3rd Meeting of the International Electrical Engineering UDC Committee and the 1st Meeting of the Telecommunications Subcommittee of the International Federation for Documentation—C. Frachebourg. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, vol. 31, pp. 30–31; Jan. 1953. In French.) Brief note of the proceedings, with proposed revised classifications for the sections on antennas and radar.

621.38/.39]:[002+026+025.45 2192

Radio and Electronic Engineering Literature—(*J. Brit. I.R.E.*, vol. 13, pp. 65–75; Jan. 1953.) A report prepared by the Technical Committee of the British Institution of Radio Engineers as a guide to the main sources of information on modern developments including abstracting services, Government publications, library facilities, patents, etc. A brief explanation is given of the working of the Patent Office Library and of the UDC (Universal Decimal Classification) system as applied to radio and electronic engineering.

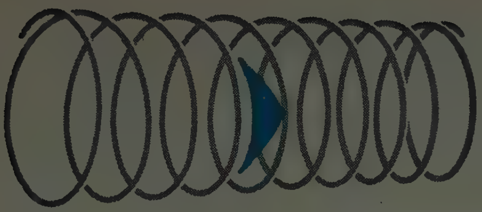
621.39(083.74/.75) 2193

Standardization and Telecommunications—R. Goret. (*Ann. Telecommun.*, vol. 8, pp. 11–18; Jan. 1953.) An account is given of the structure and operation of the commission set up in France in 1945 to study the standardization of telecommunications equipment. Specifications established by the commission are listed and discussed.

621.396 2194

Radio Progress during 1952—(Proc. I.R.E., vol. 41, pp. 452–507; May 1953.) A review including over 1000 references to world literature. Material is arranged under the headings indicated in 2404 of 1952, with additional sections on Feedback Control Systems and on Magnetics.

There's a 10-turn Helipot to meet *your* requirements



With the development of the original HELIPOT—the *first* multi-turn potentiometer—an entirely new principle of potentiometer design was introduced to the electronic industry. It made possible variable resistors combining high resolution and high precision in panel space no greater than that required for conventional single-turn potentiometers.

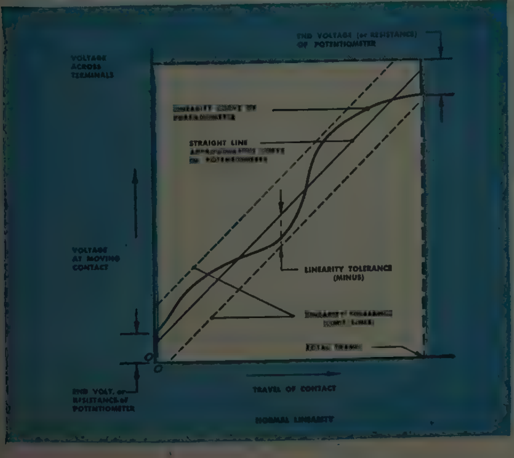
The Helipot Principle... *High resolution and precision settings require a long slide wire. But by coiling a resistance element into a helix, it is possible to gain desired resolution and precision without wasting panel space. This principle is applied in various Helipot models with slide wires ranging from 3 to 40 helical turns.*

Advantages are immediately apparent. In the case of the widely-used 10-turn Model A Helipot, for example, a 45″ long slide wire—coiled into ten helical turns—is fitted into a case 1¾″ in diameter, and 2″ in length. Another advantage of the 10-turn pot is that, when equipped with a turns-indicating RA Precision DUODIAL, slider position can be read directly as a decimal, or percentage, of total coil length traversed.

10-TURN HELIPOT MODELS—CONDENSED SPECIFICATIONS

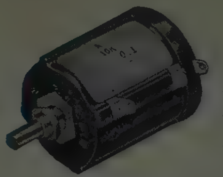
	Model A	Model AN	Model AJ
No. of turns	10	10	10
Resistance Range	10 ohms to 300,000 ohms	100 ohms to 250,000 ohms	100 ohms to 50,000 ohms
Resistance Tolerance: Standard Best	 ±5% ±1%	 ±5% ±1%	 ±5% ±3%
*Linearity Tolerance: Standard Best	 ±0.5% ±0.05% (1K ohms and above)	 ±0.5% ±0.025% (5K ohms and above)	 ±0.5% ±0.1% (above 5K ohms)
Power rating @ 40°C	5 watts	5 watts	2 watts
Mechanical Rotation	3600° +4° −0°	3600° +1° −0°	3600° +12° −0°
Electrical Rotation	3600° +4° −0°	3600° +1° −0°	3600° +12° −0°
Starting Torque	2 oz. in.	1.0±.3 oz. in.	.75 oz. in.
Running Torque	1.5 oz. in.	0.6±.3 oz. in.	.60 oz. in.
Net Weight	4 oz.	4 oz.	1 oz.

*i.e. INDEPENDENT LINEARITY. The above linearity tolerances are based on the following definition recently proposed to clarify and standardize nomenclature related to precision variable resistors. . . . "Independent linearity is the maximum deviation in per cent of the total electrical output of the actual electrical output at any point from the best straight line drawn through the output versus rotation curve. (This line shall be measured through the extent of the effective electrical angle.) The slope and position of the straight line from which the linearity deviations are measured must be so adjusted as to minimize these deviations."



10-Turn Helipot Highlights

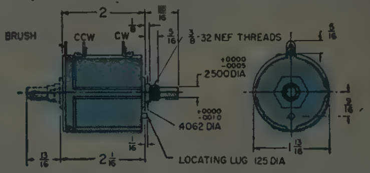
From the basic Helipot principle, model variations have been developed to meet new requirements:



Model A Helipot

the original 10-turn Helipot—provides a resolution from 12 to 14 times that of conventional single-turn potentiometers of same diameter (1¾″), linearities as close as ±0.05% in resistances as low as 1K ohms.

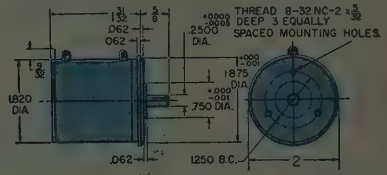
The same multi-turn principle is also available in 3 turn units (Model C), and larger-diameter units of 15 turns (Model B), 25 turns (Model D), and 40 turns (Model E)—a type for every application from 5 ohms to 1 megohm.



Model AN Helipot

an ultra-precision version of the basic 10-turn Helipot. Produced in volume to extremely close electrical and mechanical tolerances, this unit features precision ball bearings (Class 5), servo mounting lid, plus linearity tolerance as close as ±0.025% as low as 5K. A 3-turn unit (Model CN) is also available.

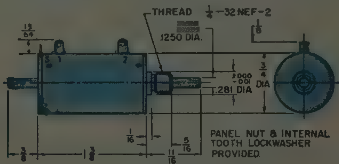
Models AN and CN are particularly recommended for precise servo-mechanism applications and represent the most advanced design and highest quality available today in the field of precision potentiometers.



Model AJ Helipot

a 10-turn *miniature* Helipot only ¾″ in diameter, weighs 1 oz., has slide wire 18″ long. Also available with servo mounting (Model AJS) and servo mounting with *ball* bearings (Model AJSP). Linearities as close as ±0.1% as low as 5K.

Designed for long life under severe operating conditions, the AJ Series is widely used where small size and weight are vital.



Design details on above units are subject to change without notice. Certified drawings available upon request.

Only Helipot is able to supply—in volume—multi-turn helical potentiometers with special features to meet your particular needs . . . Special Shafts, Extra Spot Welded Taps at any position, Ganged Assemblies (except AJ), Special Temperature Coefficients, etc. Send us your requirements!

For complete details contact your nearby Helipot representative. Or write for Data File #806

THE Helipot CORPORATION

A subsidiary of Beckman Instruments, Inc.

SOUTH PASADENA, CALIFORNIA

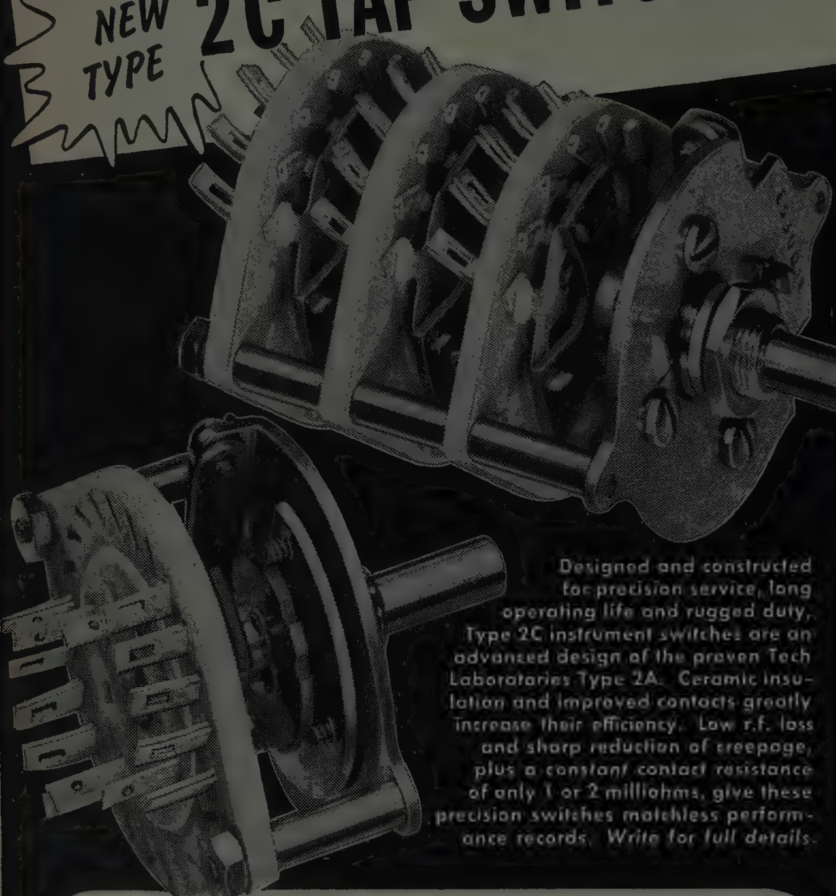
Field Offices: Boston, Rochester, New York, Philadelphia, Cleveland, Detroit, Chicago, St. Louis, Los Angeles, Seattle, Dallas, High Point, N. C. and Fort Myers, Florida. In Canada: J. S. Root, Toronto. Export Agents: Frathom Co., New York 36, New York.

Ceramic Insulated for

LOW LOSS...

MINIMUM CREEPAGE

NEW TYPE 2C TAP SWITCHES



Designed and constructed for precision service, long operating life and rugged duty, Type 2C instrument switches are an advanced design of the proven Tech Laboratories Type 2A. Ceramic insulation and improved contacts greatly increase their efficiency. Low r.f. loss and sharp reduction of creepage, plus a constant contact resistance of only 1 or 2 milliohms, give these precision switches matchless performance records. *Write for full details.*

SPECIFICATIONS

Contact resistance: 1-2 milliohms

Contact material: Silver alloy

Contact design: Laminated wiper arm, self-cleaning, shorting or non-shorting

No. of contacts: 2 to 24 single pole, shorting or non-shorting
2 to 11 double pole, shorting or non-shorting
2 to 5 triple pole, shorting or non-shorting

Spacing: 15° or 20°, shorting or non-shorting

No. of poles per deck: 1 to 4

No. of decks: According to requirements

Life: 200,000 cycles, min.

Current carrying cap.: 3 amp.

Max. operating voltage: 120 V., a.c.

Mounting: Single hole, ⅜″-32 bushing, standard length for up to ¼″ panel, special lengths to order

Size: 1¾″ dia.

Detent: Ball and spring

Weight: Approx. 1 oz. per deck

TECH LABORATORIES, INC.

Manufacturers of Precision Electrical Resistance Instruments

PALISADES PARK, NEW JERSEY

Industrial Engineering Notes*

TELEVISION

Representatives of the Federal Communication Commission recently witnessed a demonstration of compatible color television at the RCA Laboratories in Princeton, N. J. The showing was similar to the one presented last month for members of the House Interstate and Foreign Commerce Committee and was broadcast over the National Broadcasting Company's New York City station WNBT. The demonstration was held to keep FCC members informed of new developments in color. The Commission recently gave WNBT and WPPTZ, Philadelphia, permission to test the NTSC compatible color TV system during regularly scheduled operations under the following conditions: (1) the tests be made only on sustaining programs and (2) no inference be made that the color system being tested has been or will be approved by the FCC. Authorizations for the tests end July 31. . . .

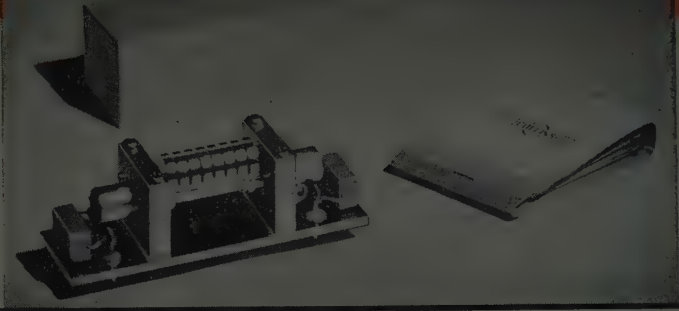
The FCC issued a reminder to broadcasters recently relative to Section 3.45 (c) of the Rules Governing Standard Broadcast Stations requirements that an application be submitted and approved by the Commission before any change is made in the physical height of an AM station antenna. A formal application, on Form 301, is required where the proposed change is more than 20 feet. In event the application is granted, the subsequent license application must be accompanied by sufficient field intensity measurements to show that the effective field has not been increased or the pattern altered, together with the exact value of the resistance added. The FCC announced that many TV applications are being filed proposing to use an AM station tower as support for the TV antenna, and in cases where an increase in height of the AM station antenna would result in increased radiation, operators have proposed that they sectionalize the tower as a method for preventing energy radiation on the AM broadcast frequency. The Commission stated, however, that it is possible to completely eliminate interference problems created by the increased antenna efficiency by dissipating enough of the power supplied to the antenna in a series resistor so as to reduce the unattenuated field to its previous value. . . .

Color television programs originated by the networks probably will be available in many local markets by the middle of 1954 and enough receivers should be on the market by that fall to permit the public generally to see and compare compatible color television with monochrome reception. This opinion was expressed recently by Dr. W. R. G. Baker, Chairman of the Nation Television System Committee and the RTMA Television Committee, when he addressed the National Association of Electrical Distributors convention in Chicago. In his talk, "Color Television—Its Status Today and a Look Into the Fu-

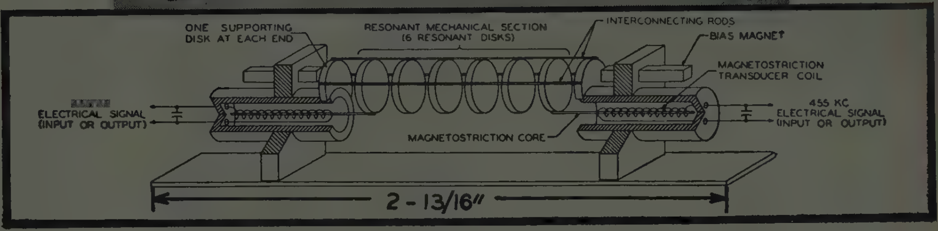
(Continued on page 70A)

* The data on which these NOTES are based were selected by permission from *Industry Reports* issues of May 22 and May 29, published by the Radio-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

MECHANICAL FILTERS FOR INDUSTRY

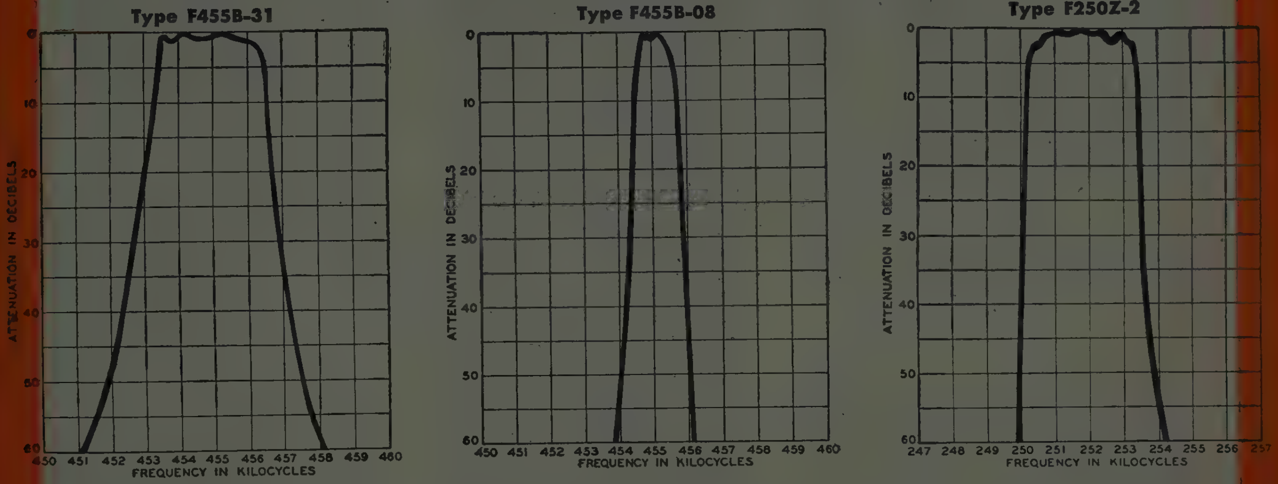


Collins Mechanical Filter with hermetically sealed shield removed. Compare size with matchbook.


ONE SUPPORTING DISK AT EACH END
RESONANT MECHANICAL SECTION (6 RESONANT DISKS)
INTERCONNECTING RODS
BIAS MAGNET
MAGNETOSTRICTION TRANSDUCER COIL
ELECTRICAL SIGNAL (INPUT OR OUTPUT)
455 KC ELECTRICAL SIGNAL (INPUT OR OUTPUT)
MAGNETOSTRICTION CORE
2 - 13/16″

Engineers's drawing of 455 kc Mechanical Filter interior.

SELECTIVITY CURVES FOR THREE TYPICAL MECHANICAL FILTERS



Since the introduction of the Collins Mechanical Filter in March 1952, the original Type F455B-31 has become available in production quantities. Other Collins Mechanical Filters, designed to operate at various bandwidths in the intermediate frequencies, are also available in engineering sample quantities. All of the Mechanical Filters listed here offer a close approach to the ideal rectangular selectivity curve. Consult the chart of available Mechanical Filters (right) for the solution to your selectivity problem.

A request on your letterhead will bring full technical specifications, price and delivery information.

Type No.	Center Frequency	Bandwidth	Terminal Arrangement
F250A-60	250 kc	6.0 kc	Solder
F250Z-1	248.7 kc	3.0 kc	Solder
F250Z-2	251.3 kc	3.0 kc	Solder
F455A-08	455 kc	800 cycle	Solder
F455A-31	455 kc	3.1 kc	Solder
F455A-60	455 kc	6.0 kc	Solder
F455B-08	455 kc	800 cycle	Plug-In
F455B-31	455 kc	3.1 kc	Plug-In
F455B-60	455 kc	6.0 kc	Plug-In
F500A-14	500 kc	1.4 kc	Solder
F500A-31	500 kc	3.1 kc	Solder
F500B-31	500 kc	3.1 kc	Plug-In
F500B-14	500 kc	1.4 kc	Plug-In

See Collins Mechanical Filters and other Collins developments at the Western Electronic Show and Convention, San Francisco Civic Auditorium, August 19, 20, 21.

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

11 W. 42nd St., NEW YORK 36 **1930 Hi-Line Drive, DALLAS 2** **2700 W. Olive Ave., BURBANK**

precision connectors by Continental

ACTUAL SIZE

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Connectors Series SM-20

afford extreme size reduction without sacrificing pin diameter . . . available in 11, 14, 20, and 34 contacts for #20 AWG wire . . . 5 amp. continuous current rating . . . Submit your special subminiature connector requirements to our engineering department.

(ACTUAL SIZE)

Continental Connectors

ELECTRONIC SALES DIVISION

DeJUR-AMSCO CORPORATION

Write Dept. P-8, DeJur-Amsco Corporation, 45-01 Northern Blvd., Long Island City 1, N.Y.
West Coast: 405 North Maple Drive, Beverly Hills, Calif.

(Continued from page 66A)

ture," Dr. Baker emphasized that his remarks "represent just one man's considered opinion. It is based primarily on my work with the NTSC covering a period of two years of fairly exhaustive deliberation and study." He pointed out that the NTSC system "is now undergoing final and extensive field tests" and "it now appears possible that the NTSC can conclude its technical work by September." He said it might be necessary to allow 45 to 60 days for the complete organization of all data to be presented to the FCC, and continued: "If such a schedule can be met we can assume that the NTSC would petition the FCC for a hearing in October or early November. This could mean that the hearing could be scheduled early in 1954 and the NTSC system could be approved by March 1, 1954. . . .

Federal Personnel

Cmdr. P. K. Beisel has been named Chief of the Electronics Branch, Production Division, Office of Navy Material. He succeeds Capt. W. I. Bull who was transferred to the Bureau of Ships as Director of the Electronic Ship and Amphibious Division.

Mobilization

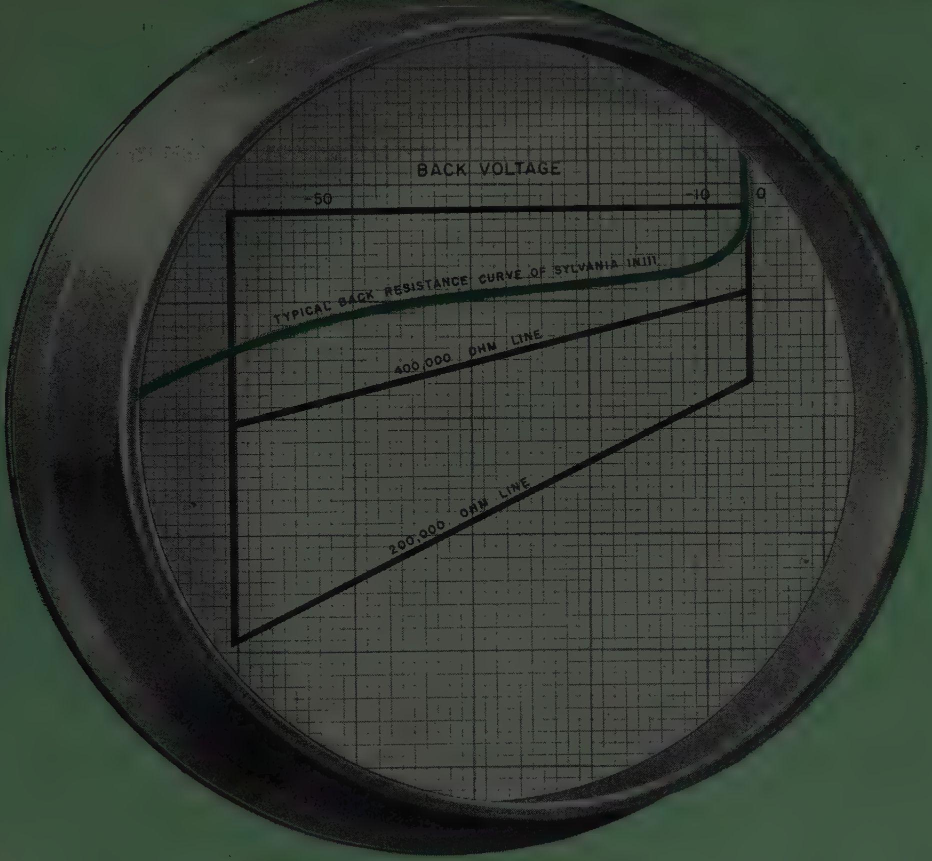
The Atomic Energy Commission recently released 20 additional patents, including five in the electronics field. Nonexclusive, royalty-free licenses on the listed patients will be granted by the Commission as part of its program to make nonsecret technological information available for use by industry. Applicants for licenses should apply to the Chief, Patent Branch, Office of the General Counsel, U. S. Atomic Energy Commission, Washington 25, D. C., identifying the subject matter by patent number and title. The following patents of interest to the electronics industry were released: High Frequency Shunt, 2,634,307; Pocket Radiation Meter, 2,634,-374; Pulse Equalizer, 2,636,118; Ion Source Unit, 2,636,990, and Fast Counting Circuit, 2,636,993.

FCC Actions

The FCC issued recently a Notice of Proposed Rule Making, looking toward relaxing existing requirements relating to the assignment of Class II, III and IV stations in the standard broadcast band by amending its "10 per cent rule." The proposed amendment would permit a Class II, III or IV station to be assigned to a channel available for such class when a need is shown, even though interference will be received within its normally protected contour, provided (1) no objectionable interference will be caused by it to existing stations or that if interference will be caused the need for the proposed service outweighs the need for the service which will be lost by reason of such interference (thus incorporating into the Standards the same provision now contained in Section 3.24(b) of the Commission's rules), and (2) primary service will be provided to the

(Continued on page 72A)

Sylvania Computer Crystal [illegible]



1N111
1N112
1N113
1N114
1N115

All Dynamically Tested at 55° C.
For High Back Resistance and Stability

Sylvania Types 1N111, 1N112, 1N113, 1N114 and 1N115 were designed specifically for computer use. All Sylvania's Computer Diodes are tested at raised temperatures simulating actual operating conditions. To insure maximum stability and life, all units are tested for evidence of drift and hysteresis. Each diode is hermetically sealed in glass and is designed so that it may conveniently be soldered or clipped into a circuit.

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LORD TEMPROOF MOUNTINGS ON COLLINS 618S-1 TRANSCEIVER

SENSITIVE electronic equipment for airline transmitting and receiving must give continuously accurate results. For instance, note this "inside" view of the Collins Transceiver, mounted on Lord Temproof Mountings which isolate it from vibration and shock. Lord Temproof Mountings function efficiently throughout operational ranges of temperature from —80° to +250°F. The Collins Transceiver with automatically tuned elements for maximum flexibility and high power output delivers maximum performance in any weather, completely protected from vibration, shock and excessive equipment motion at resonant frequencies by Lord Temproof Mountings.

May we give you further details on this Lord application or help you solve your specific mounting requirement?

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DALLAS, TEXAS
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PHILADELPHIA 7, PENNSYLVANIA
725 Widener Building

DAYTON 2, OHIO
410 West First Street

DETROIT 2, MICHIGAN
311 Curtis Building

NEW YORK 16, NEW YORK
280 Madison Avenue

CHICAGO 11, ILLINOIS
520 N. Michigan Ave.

CLEVELAND 15, OHIO
Room 811 Hanna Building

LORD MANUFACTURING COMPANY • ERIE, PA.

LORD BONDED RUBBER
Headquarters for VIBRATION CONTROL

Industrial Engineering Notes

(Continued from page 70A)

community in which the station is to be located. In addition, the assignment must meet one of the following conditions: (a) provide the city with its first such local stations; (b) provide the first primary service to 25 per cent or more of its primary service area; or (c) interference does not affect more than 10 per cent of the population in the station's normally protected primary service area. . . .

The Commission recently also adopted a Second Further Notice of Proposed Rule Making which lists specific dates for changing from certain old to certain new frequencies by ship and coast stations using maritime mobile radio-telephone frequencies in the band 4,000–18,000 kc for coastal areas of the United States and Hawaii. This further supplements the Commission Report and Order of May 6, 1953, to implement the Geneva agreement. . . .

The Commission issued a Notice of Proposed Rule Making recently to amend Section 12.91(b) of Part 12, "Rules Governing Amateur Radio Service," to permit maritime mobile amateur radio stations to operate on frequencies in the band 21,000–21,450 kc when outside the continental limits of the United States, its territories and possessions. Comments may be filed on or before August 1, 1953. Another Notice of Proposed Rule Making, issued by the Commission recently, looks toward amending Parts 10, 11, and 16 of the Rules to establish certain new provisions regarding operational fixed stations operating on frequencies above 89-meg. It has been considered appropriate, FCC said, to standardize on the specific frequencies to be assigned to the bands above 890 meg, or to establish uniform technical standards for the stations operating on those bands. The FCC, therefore, sees the need for some degree of standardization in the bands 952–960 meg. where the majority of operation is taking place. . . .

The FCC recently finalized a previous Notice of Proposed Rule Making by amending Part 8 of the Rules Governing Stations on Shipboard in the Maritime Service so as to delete the frequency 8,830 kc from the assignable frequencies for ship telephone use, and to make 8,262.3 kc and 8,219.7 kc available for ship telephone stations, effective June 1, 1953.

SECTION MEETINGS

AKRON

"Frequency Modulated Radar," by R. C. Sanders, Jr., Sanders Associates; Election of officers; May 19, 1953.

BALTIMORE

"Telemetering Instruments, Systems and Techniques," by M. V. Kiebert, Jr., Consulting Engineer, election of officers; June 10, 1953.

(Continued on page 74A)

Telecomputing announces . . .

THE TELEDUCER

Automatically Records Voltage

In Decimal Digits

With 0.1% Accuracy

It offers these important advantages:

Provides 0.1% accuracy (1,000 counts full scale).

Provides for minimum full-scale input of 20 millivolts (20 microvolts per count).

Requires only 0.8 second or less for balancing.

Uses a simple bridge-balancing circuit.

Reads low voltage without DC Amplification.

Does not hunt or oscillate.

Presents visual reading of voltage on neon indicator bank.

Permits tabulated digital output when used with Program Unit and Electric Typewriter. (Digital recording in punched IBM cards or teletype tape available when used with suitable program units.)

You are invited to visit Telecomputing's exhibit, Booths 1722 and 1723, at the Western Electronics Show in the Civic Auditorium in San Francisco, August 19-21.

TELEDUCER Technical Bulletin 104 and Technical Bulletins 106 and 107, giving specifications on the Program Unit and Electric typewriter, will be mailed you upon request. Coupon below is for your convenience.

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Telecomputing Corporation, Burbank, California

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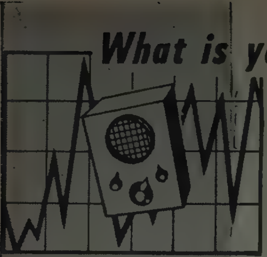
BURBANK, CALIFORNIA

What is your Delay or Regulating Problem?



For the most effective solution use the
SIMPLEST, MOST COMPACT
MOST ECONOMICAL
HERMETICALLY SEALED

AMPERITE
THERMOSTATIC DELAY RELAYS

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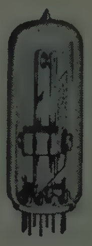
Provide delays ranging from 2 to 120 seconds.
- Actuated by a heater, they operate on A.C., D.C., or Pulsating Current.
- *Hermetically sealed.* Not affected by altitude, moisture, or other climate changes.
- Circuits: *SPST only*—normally open or normally closed.

Amperite Thermostatic Delay Relays are compensated for ambient temperature changes from −55° to +70°C. Heaters consume approximately 2 W. and may be operated continuously. The units are *most compact, rugged, explosion-proof, long-lived,* and—*very inexpensive!*



MINIATURE

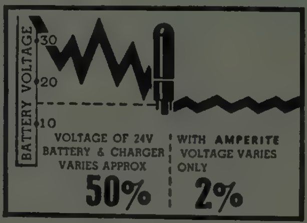
TYPES: Standard Radio Octal, and 9-Pin Miniature.
PROBLEM? Send for Bulletin No. TR-81

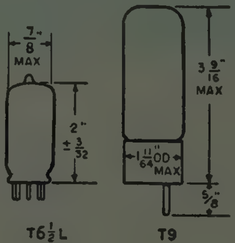
BALLAST-REGULATORS

AMPERITE
REGULATOR

T9 BULB

- Amperite Regulators are designed to keep the current in a circuit *automatically regulated* at a definite value (for example, 0.5 amp).
- For currents of 60 ma. to 5 amps. Operates on *A.C., D.C., or Pulsating Current.*
- Hermetically sealed, light, compact, and most inexpensive.





Maximum Wattage Dissipation: T6½L—5W. T9—10W.

Amperite Regulators are the simplest, most effective method for obtaining *automatic regulation* of current or voltage. *Hermetically sealed,* they are not affected by changes in altitude, ambient temperature (−55° to +90°C), or humidity. Rugged; no moving parts; changed as easily as a radio tube.

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SECTION MEETINGS

(*Continued from page 72A*)

BINGHAMTON

"Ground Combat Operation in Korea from an Electronic Viewpoint," by W. J. Nelson, General Electric; and "Ire Organization and Future Plans," by C. A. Priest, IRE Regional Director; May 24, 1953.

BOSTON

"Some New Results in Time and Frequency-Domain Synthesis," by E. A. Guillemin, Faculty, Mass. Institute of Technology; and "Low-Brow Applications of High-Brow Circuit Theory," by R. E. Scott, Faculty, Mass. Institute of Technology; April 23, 1953.

BUFFALO-NIAGARA

"Transistor Circuits and Television Applications," by G. C. Sziklai, RCA Labs.; May 20, 1953.

CEDAR RAPIDS

"Radio Astronomy and the 50 foot Radio Telescope," by Dr. J. P. Hagen, Naval Research Laboratory; film: "Stepping Along with Television"; April 22, 1953.

"The Kerr Effect and its Utilization as a Camera Shutter," by Dominic Aversa, "A Method of Determining the Value of the Elementary Charge," by Gaylord Treu and "Semi-Conductors Become of Age," by J. D. Miller, all students, University of Iowa; "Status of TV in Cedar Rapids," by George Hixenbaugh, WMT; May 20, 1953.

CHICAGO

"The Self-Transistor Effect in Germanium Diodes," by B. R. Gossick, Purdue University; election of officers; May 15, 1953.

CLEVELAND

"Measuring Information," by Dr. David Slepian, Bell Telephone Labs.; election of officers; May 28, 1953.

COLUMBUS

"Measuring Information," by Dr. D. Slepian, Bell Telephone Labs.; May 27, 1953.

CONNECTICUT VALLEY

Demonstration on colored television by C. N. Hoyler, R.C.A.; March 11, 1953.

New England Radio Engineering meeting; April 11, 1953.

"Transistors and Transistor Circuits," by Prof. H. J. Reich and Prof. J. May, Yale University; election of officers; May 21, 1953.

DALLAS-FORT WORTH

Installation of officers; June 5, 1953.

DENVER

"Color and Color Television," by Dr. R. C. Webb, Denver Research Institute; election of officers; June 5, 1953.

DES MOINES-AMES

"Electrical Analogs for Heating Systems," by E. E. Jones, Minneapolis Honeywell Co.; May 28, 1953.

DETROIT

Student papers: "The Miracle Fuze," by T. E. Burke, Michigan State College, "Electronic Indicator for the Analytical Beam Balance," by C. C. Mosher, Michigan State College and "Theory and Application of Negative Impedance," by Der-Shyang Ling and Richard Putansu, University of Detroit; May 22, 1953.

ELMIRA-CORNING

"Cathode Ray Tube Engineering and Manufacturing Facilities of Westinghouse," by C. R. Potter, W. J. Knochel, J. A. Hall, H. F. Pully and J. A. Rima; election of officers; May 18, 1953.

EVANSVILLE-OWENSBORO

"What's Ahead in Engineering," by E. S. Lee, Editor, General Electric Review; election of officers; May 13, 1953.

"Color Television," by J. N. Phillips, General Electric Company; June 10, 1953.

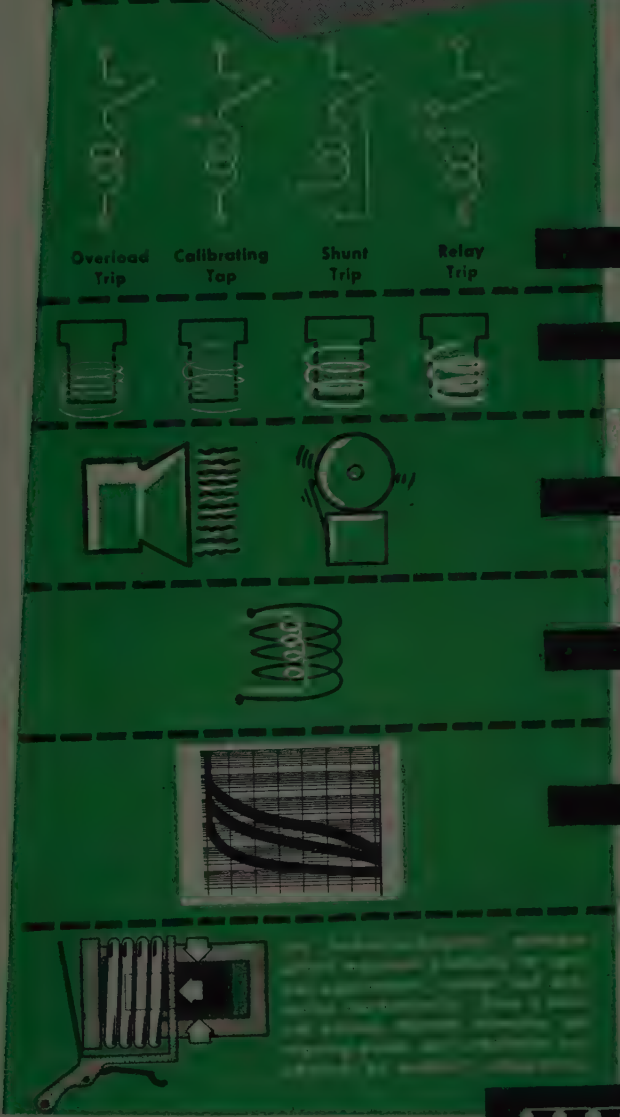
(*Continued on page 76A*)

Need Something Special in a Circuit Breaker?

Perhaps the problem of protection or control of your equipment can be solved by a HEINEMANN Circuit Breaker.

Operating on a hydraulic-magnetic, solenoid principle . . . with or without inverse time delay . . . HEINEMANN Circuit Breakers offer almost unlimited possibilities in fulfilling standard or special functions.

Your requests for special information will receive prompt attention.



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Close tolerances of protection may be obtained by precise, fractional rating between 10 milliamperes and 100 amperes.

ALARM CONTACTS

Alarm contacts are available on certain models to permit visual or audible signal on tripping of circuit breaker.

DUAL RATINGS

A single circuit breaker may be furnished with two coils for operation on different currents.

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Time vs. per cent overload response curves may be selected to match protection characteristics to the requirements of your product.

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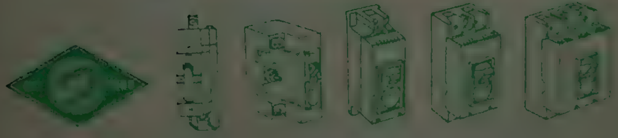
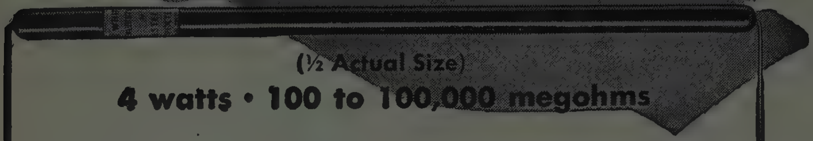


HEINEMANN Circuit Breakers...One, two and three pole...10 milliamps to 100 amperes

The new S.S.WHITE 80X
HIGH VOLTAGE RESISTOR



(½ Actual Size)
4 watts • 100 to 100,000 megohms

Developed for use as potential dividers in high voltage electrostatic generators, S.S.White 80X Resistors have many characteristics—particularly negative temperature and voltage coefficients—which make them suitable for other high voltage applications.

They are constructed of a mixture of conducting material and binder made by a process which assures adequate mechanical strength and durability. This material is non-hygroscopic and, therefore, moisture-resistant. The resistors are also coated with General Electric Dri-film which further protects them against humidity and also stabilizes the resistors.

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SECTION MEETINGS

(Continued from page 74A)

FORT WAYNE

"Telemetering Instruments and Systems," by M. V. Kiebert, Jr., Eclipse-Pioneer Division—Bendix; election of officers; May 20, 1953.

HOUSTON

"Transistors," by Lester Hill, Student, Texas A & M College; "A Direct Reading Phase Meter," by William Heard, Student, University of Houston; "The Use of Switching Algebra in Contact Network Design," by George McDaniel, Student, Rice University; election of officers; April 28, 1953.

LOS ANGELES

"CBS Television City," by Les Bowman, CBS; "Operation Greenhouse" (AEC film prefaced by brief description of project) by D. L. Narver, Jr., Eniwetok Proving Ground—JTF-3, Holmes and Narver, Inc.; June 2, 1953.

LOUISVILLE

Election of officers; June 11, 1953.

MILWAUKEE

"Germanium Power Rectifiers," by A. C. Sheckler, General Electric Company; May 21, 1953.

NEW YORK

"Beam-Hugging Plates for Unlimited Cathode Ray Deflection," by H. E. Kallman, Research Consultant and "New Thyratron Control Characteristic by Ion Deflection," by G. W. Baker, Kip Electronics Corporation; May 6, 1953.

"Electronic Conduction in Solids," by K. K. Darrow, Bell Telephone Labs.; election of officers; June 3, 1953.

OMAHA-LINCOLN

"Reproduction of Binaural Recordings," by Clyde Scriver, Magnacorder Co. of Chicago; May 7, 1953.

OTTAWA

Field Trip—microwave relay system, Bell Telephone Company of Canada; May 21, 1953.

PHOENIX

"Reliability of Electronic Equipment," by Dr. Clarence Gates, Jet Propulsion Group; election of officers; May 22, 1953.

PITTSBURGH

"Quality Assurance by Statistical Methods," by D. E. Smith, Faculty, Pittsburgh Technical Institute; June 8, 1953.

PRINCETON

"The VOR in Airline Operations," by E. A. Post, United Air Lines; election of officers; May 14, 1953.

ROCHESTER

"The Transistor-Vacuum Tube of Tomorrow," by W. J. Stolze, Stromberg-Carlson Co." election of officers; May 21, 1953.

SAN ANTONIO

"General Human Communications," by G. P. Harnwell, University of Pennsylvania; May 14, 1953.

SAN FRANCISCO

"Color TV," by Axel Jensen, Bell Telephone Laboratories; May 12, 1953.

Student contest: "The Design of an Alerting Device for Conelrad Operation," by Frank Glass; "An Engineering Approach to Low Priced Audio Circuitry," by G. S. Bahrs, "A Complex Frequency Generator," by Yale Barkan and "Piezoelectricity in Electrets," by William Duffy; May 20, 1953.

(Continued on page 78A)

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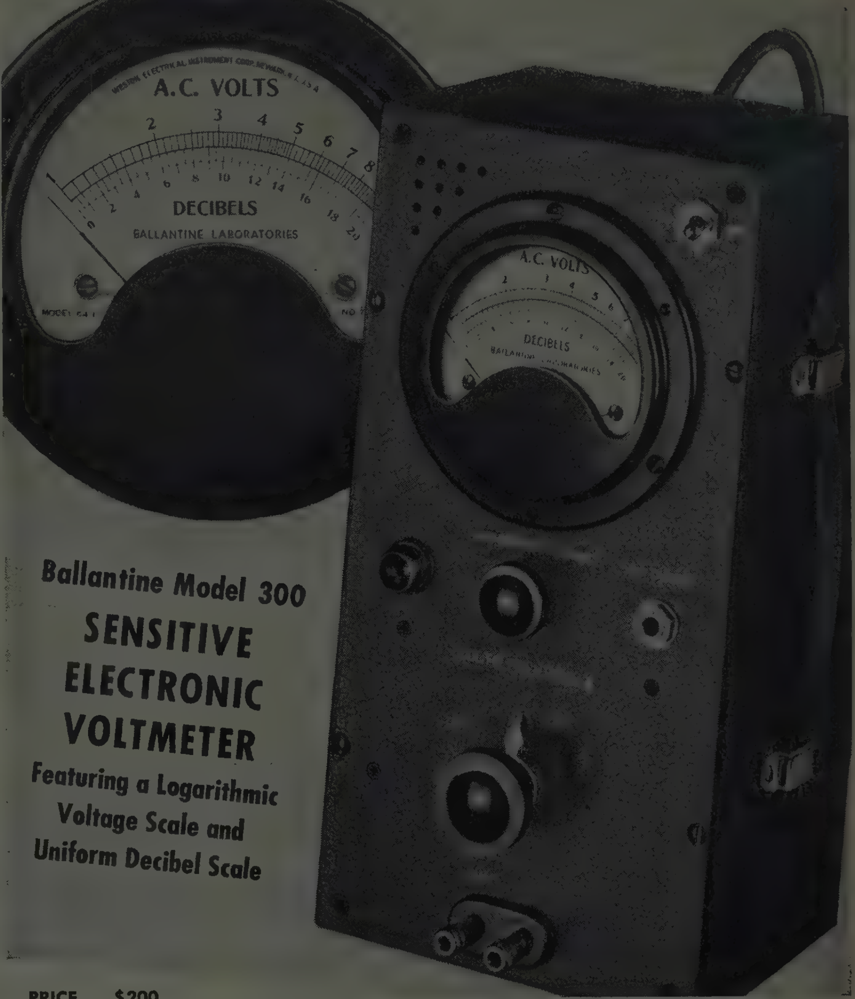

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SECTION MEETINGS

(Continued from page 76A)

SCHENECTADY

"High Power Klystron and Helical Antenna for UHF Telecasting," by H. B. Faucher; General Electric Company; April 13, 1953.

SYRACUSE

"Information Theory," by Dr. Stanford Goldman, Syracuse University; April 16, 1953.

"Ferrites," by Dr. H. C. Rothenberg, General Electric Co.; election of officers; presentation of student awards; May 7, 1953.

TORONTO

"CBC Television," by J. A. Ouimet, Canadian Broadcasting Corp.; April 27, 1953.

TULSA

Business meeting; election of officers; May 27, 1953.

TWIN CITIES

"Advancements in Design of Wide Band Amplifiers," by Dr. R. M. Pettit, Stanford Research Foundation; March 18, 1953.

"Miniaturized Components for Transistor Electronics," by P. S. Darnell, Bell Telephone Labs.; May 6, 1953.

"High Speed Automatic Storing and Composing of Teletype Messages," by J. L. Hill, Remington Rand, Inc.; May 21, 1953.

Inspection Trip of Toll Dialing Machine; election of officers; June 3, 1953.

VANCOUVER

"Stereophonic Sound," by J. Hauser, Ampex Corp.; April 20, 1953.

"Digital Computers," by Mr. Forsman, International Business Machines; May 25, 1953.

WASHINGTON, D. C.

"Broadcast Color Television, the Problems and the Prospects," by A. V. Loughren and C. J. Hirsch, Hazeltine Electronics Corp.; May 11, 1953.

"Our Expanding Technology," by Dr. J. W. McRae, President, IRE; election of officers; June 8, 1953.

SUBSECTIONS

ITHACA

"Military Electronics in Korea," by J. W. Nelson, General Electric Company; election of officers; March 18, 1953.

LANCASTER

"Principles of the NTSC Compatible Color System," by C. J. Hirsch, Hazeltine Corporation; election of officers; May 13, 1953.

MID-HUDSON

"Projection Television," by J. Deerhake, General Electric Company; election of officers; April 21, 1953.

"Conservation," by Prof. John George, Vassar College; May 21, 1953.

MONMOUTH

"A Maze Solving Mouse," by Claude Shannon, Bell Telephone Labs.; election of officers; May 20, 1953.

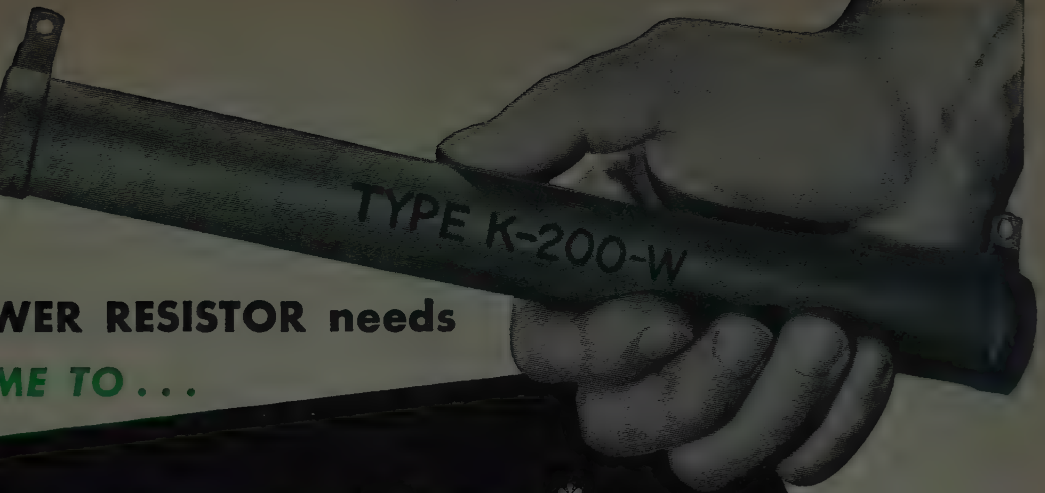
NORTHERN NEW JERSEY

"An Ultrasonic Machine Tool," by N. B. Nichols and Neil Clark, Raytheon Manufacturing Company; election of officers; May 13, 1953.

ORANGE BELT

"California Institute of Technology Synchrotron," by Dr. R. V. Langmuir, California Institute of Technology; June 17, 1953.

(Continued on page 80A)



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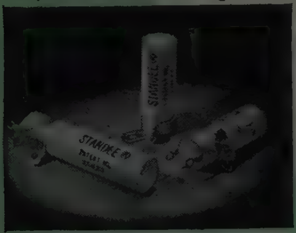

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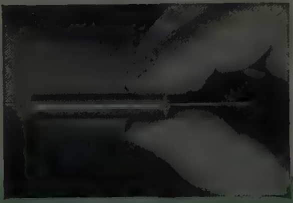

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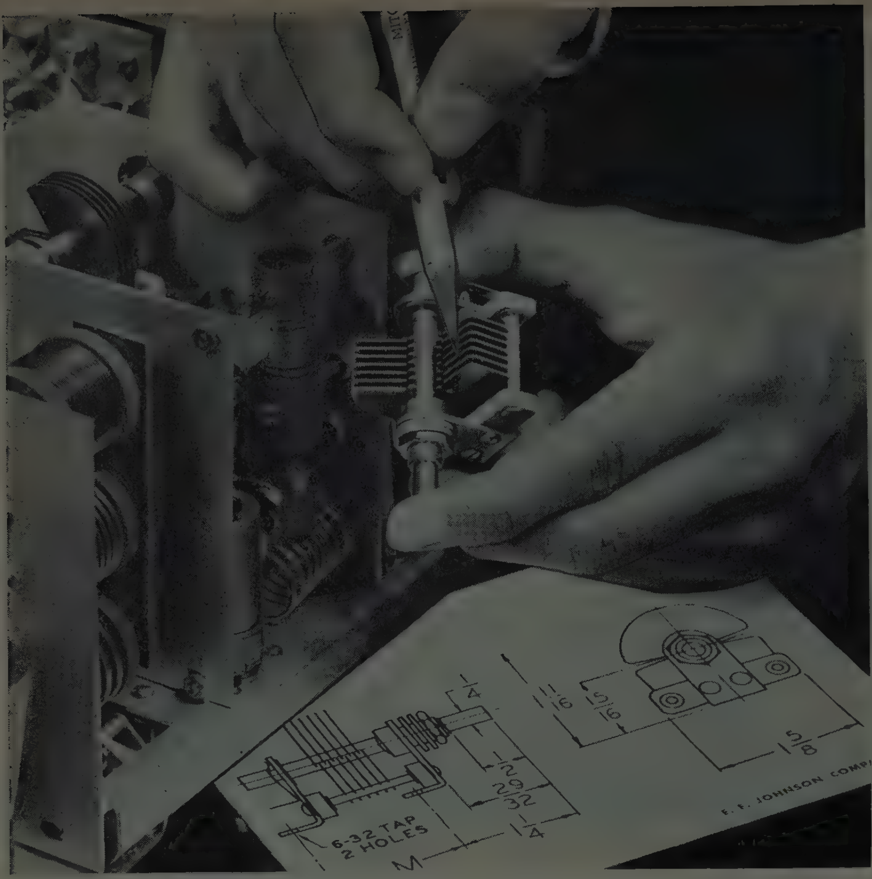

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SECTION MEETINGS

(Continued from page 78A)

Palo Alto

"Operations Analysis—the New Science of Evaluating the Effectiveness of Military Operations," by T. E. Oberbeck, W. P. Cunningham, C. C. Torrance, Faculty, U. S. Naval Postgraduate School; May 11, 1953.

"The Role of International Broadcasting in the Cold War," by James Day, Radio Free Asia; election of officers; June 5, 1953.

Rome

"Transistors, Today and Tomorrow," by H. L. Owens, Evans Signal Laboratory; May 25, 1953.

Wichita

"Practical Applications of Transistors," by L. E. Flory, R.C.A.; May 20, 1953.

Winnipeg

"Electronics of the Brain," by Dr. M. G. Saunders, Manitoba Medical College; May 14, 1953.

"The Role of Electronics in Radio Therapeuty," by Dr. H. Blondal, Manitoba Medical College; June 3, 1953.

Annual meeting; election of officers; June 17, 1953.

STUDENT BRANCH MEETINGS

Brooklyn Polytechnic Institute (Day Div.), IRE Branch

"The Transistor and Some of Its Applications," by G. A. Hildebrand, Radio Corporation of America; March 10, 1953.

"The Theory of Junction Transistors and Applications," by G. Raisbeck, Bell Telephone Labs.; March 21, 1953.

Election of officers; April 21, 1953.

Brown University, IRE-AIEE Branch

"Research in Noise Analysis," by Dr. M. F. Gordon, Faculty, Brown University; election of officers; April 9, 1953.

Film, "Electrical Proving Ground"; May 6, 1953.

University of California, IRE-AIEE Branch

General meeting; March 3, 1953.
Field trip; March 6, 1953.
Executive meeting; March 17, 1953.
Executive meeting; March 24, 1953.
General meeting; May 12, 1953.
Nomination of officers for Fall semester; May 19, 1953.
Executive meeting; May 26, 1953.

California State Polytechnic College, IRE Branch

"Prospects of Colored Television," by A. G. Jensen, Bell Telephone Labs.; May 7, 1953.

University of Dayton, IRE-AIEE Branch

"Electrical Meters," by Henry Berring, Weston Electrical Company; April 28, 1953.

Election of officers; May 5, 1953.

University of Detroit, IRE-AIEE Branch

Election of officers and technical papers: "Magnetic Amplifiers," by Ken Popis; "Ductile Bismuth Wire," by R. Klepazyk and C. Stout; "Induction Heating of Conducting and Non-Conducting Materials," by J. Lienhard and T. Lienhard; "Theory and

(Continued on page 82A)

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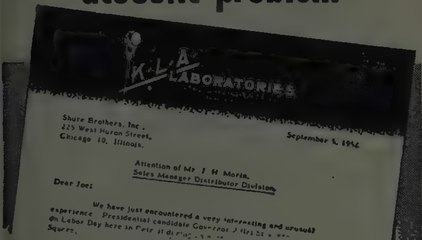

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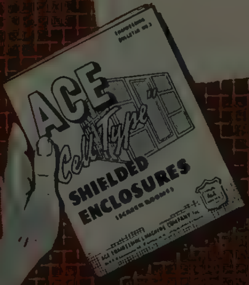

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STODDART AIRCRAFT RADIO Co., Inc.

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STUDENT BRANCH MEETINGS

(Continued from page 80A)

Application of Negative Impedance," by D. S. Ling and R. Putansu; "Corona Discharge of Transmission Lines," by J. Papp and "Theory and Operation of Germanium Diodes and Triodes," by H. McKenna; April 23, 1953.

UNIVERSITY OF FLORIDA, IRE-AIEE BRANCH

Election of officers; April 28, 1953.

General meeting; May 12, 1953.

GEORGIA INSTITUTE OF TECHNOLOGY, IRE BRANCH

"IBM Computers," by H. D. Ivey, Georgia Tech Engineering Experiment Station; color film on IBM Computers; May 26, 1953.

ILLINOIS INSTITUTE OF TECHNOLOGY, IRE BRANCH

"The Basic Elements of Electronic Computers," by David Rubinfien, Armour Research Foundation; May 11, 1953.

"Electronic Computers and Devices," by Mr. Moan, IBM Company; May 13, 1953.

IOWA STATE COLLEGE, IRE-AIEE BRANCH

"Inside Digital Computers," by F. C. Mullaney, Engineering Research Associates; March 4, 1953.

"Circuit Applications of Transistors," by T. R. Finch, Bell Telephone Labs.; March 9, 1953.

"Engineering the Automatic Factory," by E. S. Sampson of General Electric; April 8, 1953.

"Alumni Needs Growing Out of Alumni Work," by Wallace Barron; May 1, 1953.

Election of officers; May 20, 1953.

STATE UNIVERSITY OF IOWA, IRE BRANCH

Field Trip to WSUI Transmitter; May 20, 1953.

Election of officers; May 27, 1953.

LAFAYETTE COLLEGE, IRE-AIEE BRANCH

"Phases of Sales in Utilities," by J. F. Roberts, Philadelphia Electric; March 25, 1953.

"Traveling Wave Tubes," by Roger White, Federal Telecommunications Labs.; April 14, 1953.

MANHATTAN COLLEGE, IRE-AIEE BRANCH

Film: "Laying of 138 KV Cable Under the Narrows." Speaker: Mr. Shaughnessey, Consolidated Edison; May 6, 1953.

Field trip to Sewaren Generating Plant; May 14, 1953.

Business meeting and election of officers; May 22, 1953.

MASSACHUSETTS INSTITUTE OF TECHNOLOGY, IRE-AIEE BRANCH

"Desk Fax"—A facsimile telegram transmission device; March 17, 1953.

"Modern Communication Techniques," by Mr. Osborne, American Tel. and Tel. Co.; April 16, 1953.

"The Mass. Institute of Technology Telephone System," by Prof. C. E. Tucker, Faculty, Mass. Institute of Technology; April 28, 1953.

"An Outline of the Changes in the E.E. Dept. of M.I.T.," by Prof. G. S. Brown, Faculty, Mass. Institute of Technology; May 5, 1953.

UNIVERSITY OF MICHIGAN, IRE-AIEE BRANCH

Election of officers and "Television Engineering," by Clyde Hallmark, Magnavox Company; May 6, 1953.

"Opportunities in Nuclear Engineering," by Dr. H. J. Gomberg, University of Michigan Phoenix Project; May 20, 1953.

(Continued on page 86A)

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DUMONT

ALLEN B. DU MONT LABORATORIES, INC.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 48A)

Nuclear Film Badge Service

A complete film badge service for improved personnel protection has been announced by the **Nuclear Instrument & Chemical Corp.,** 229 W. Erie St., Chicago 10, Ill. This "Nuclibadge" service consists of a film badge and holder designed for accurate interpretation of accumulated X and gamma radiation over a range of 50 mr to 500 r with energies from 50 Kev to 20 Mev.

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The badge number, the date the badge is to be worn, and the wearer's name will appear on both the outside of the film packet and on the film emulsion. Control films to monitor radiation accumulated in transit will be provided without extra charge with service of three or more badges per week. After processing, all films will be returned to the user along with an interpretation and a written report.

(Continued on page 91A)

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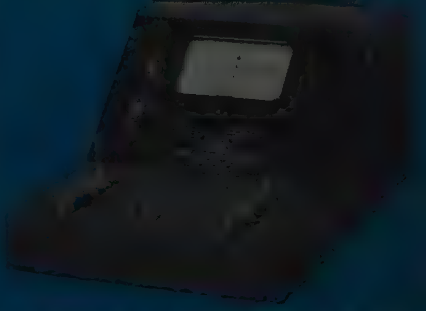

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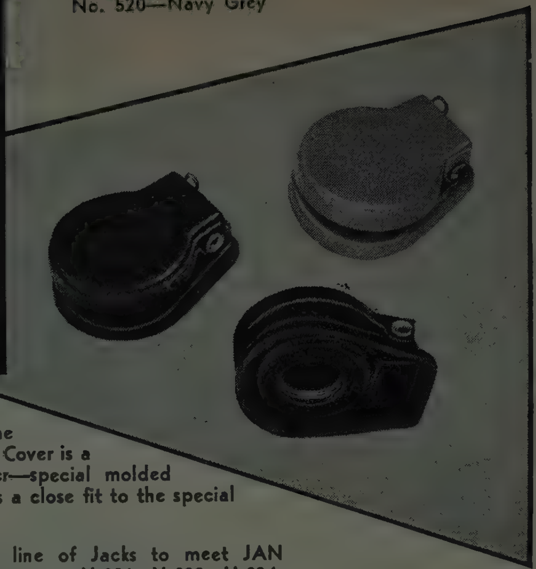

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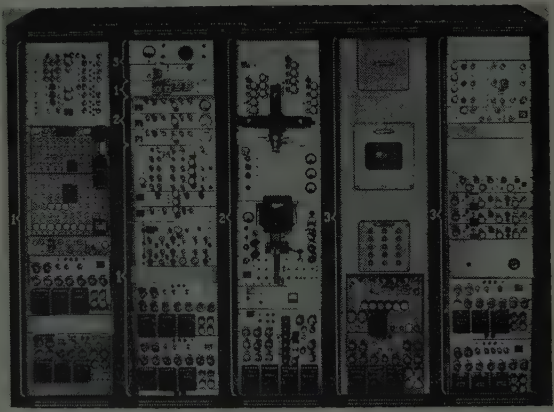

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GROUP 2— Additional Equipment includes flying spot picture generator and gamma amplifiers; permits encoding of 35 mm color slides.

GROUP 3 — Complete Equipment contains complete decoding and tri-color display, video amplifiers and sync separator, picture and sound transmitter.

COLOR TV TELECHROME INCORPORATED

THROUGH ADVANCED RESEARCH & DEVELOPMENT

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TELEPHONE: AMITYVILLE 4-4446

STUDENT BRANCH MEETINGS

(Continued from page 82A)

UNIVERSITY OF MINNESOTA, IRE-AIEE BRANCH

"Analog and Digital Computers," by P. N. Hess and Mr. Fenne, Research Associates, Univ. of Minnesota; Election of officers; May 4, 1953.

"Microwave Spectroscopy," by H. Crawford; June 9, 1953.

COLLEGE OF THE CITY OF NEW YORK, IRE BRANCH

Election of officers; May 28, 1953.

NEW YORK UNIVERSITY (EVE. DIV.), IRE-AIEE BRANCH

"Circuit Break Construction and Testing Technique," by F. H. Stoppelmann, Westinghouse Electric Corp. Slides: 3d views of circuit breaker construction; May 8, 1953.

Election of officers; June 12, 1953.

OHIO STATE UNIVERSITY, IRE-AIEE BRANCH

"Electrical Indicating Instruments," by Henry Berring, Weston Electric Instrument Corp.; Election of officers; May 4, 1953.

Films, "The Loose Nut" and "Amazon Awakens"; May 21, 1953.

OKLAHOMA AGRICULTURE & MECHANICAL COLLEGE, IRE-AIEE BRANCH

"Problems in Short Line TV Signal Transmission," by Wallace Fullerton, Southwestern Bell Telephone Co.; May 4, 1953.

UNIVERSITY OF PITTSBURGH, IRE BRANCH

"Educational Television in Pittsburgh," by V. W. Adams, University of Pittsburgh; May 14, 1953.

SAN DIEGO STATE COLLEGE, IRE BRANCH

Election of officers; May 21, 1953.

SAN JOSE STATE COLLEGE, IRE BRANCH

"Development of the Robot Can Ejector," by L. E. Worcester, Illumitronic Engr. Co.; May 22, 1953.

SEATTLE UNIVERSITY, IRE BRANCH

Business meeting; April 18, 1953.

UNIVERSITY OF SOUTHERN CALIFORNIA, IRE-AIEE BRANCH

Election of officers; May 25, 1953.

UNIVERSITY OF UTAH, IRE-AIEE BRANCH

General meeting; April 9, 1953.
General meeting; April 16, 1953.
General meeting; April 23, 1953.
General meeting; May 14, 1953.
General meeting; May 21, 1953.
General meeting; May 28, 1953.

UNIVERSITY OF WASHINGTON, IRE-AIEE BRANCH

Grand Coulee Field Trip; May 2 and 3, 1953.
Business meeting; May 12, 1953.

UNIVERSITY OF WISCONSIN, IRE BRANCH

"Electrical Indicating Instruments," by Henry Berring, Weston Instrument Corp.; March 26, 1953.

Election of officers and student paper competition: "High Speed Digital Computing," by Wayne Olson, "Automatic Control and Modern Society," by Ted Martin and "Mechanical Aspects of Overhead Electrical Construction," by Tom Benedict; May 7, 1953.

MEMBERSHIP

The following transfers and admissions were approved to be effective as of August 1, 1953:

Transfer to Senior Member

Alba, C. J., 1113 N. Pitt, Alexandria, Va.
Anderson, D. A., 159 Sunnyside Ave., Lakeside, Montreal, Que., Canada
Aron, M., Radio Receptor Co., Inc., 84 N. Ninth St., Brooklyn, N. Y.
Atchison, F. S., 6197 Palm Ave., Riverside, Calif.
Bachman, W. J., 17 Beechwood Rd., Florham Park, N. J.
Bauer, J. A., 2411 Browning Rd., Pennsauken 8, N. J.
DeYoung, J. A., 260 C Main St., Medford 55, Mass.
Elias, L. G., 99 Elm St., Danbury, Conn.
Femmer, M. E., R.F.D. 2, Hopewell Junction, N. Y.
Foley, G. M., c/o Leeds & Northrup Co., Philadelphia 44, Pa.
Fox, P. E., Hillis Ter., R.F.D. 3, Wappingers Falls, N. Y.
Hanna, F. M., PSC Applied Research, Ltd., 1450 O'Connor Dr., Toronto 13, Ont., Canada
Henderson, V. L., Department of Applied Physics, University of Toronto, Toronto, Ont., Canada
Henthorn, J. E., Jr., 4518 Pen Lucy Rd., Baltimore 29, Md.
Herrmann, G. M., 26 Front St., Schenectady 5, N. Y.
Johnson, N., 1602 Powdermill La., Wynnewood, Pa.
Jordan, J. F., 501 Miami Ave., Terrace Park, Ohio
Long, M. L., 457 Madison Ave., Hatboro, Pa.
Lusted, L. B., Department of Radiology, University of California Hospital, San Francisco 22, Calif.
Marx, E. A., 350 Fifth Ave., New York 1, N. Y.
Montgomery A. P., 5515 Wissahickon Ave., Philadelphia 44, Pa.
Nagel, G. W., 234 Cascade Rd., Pittsburgh 21, Pa.
Ober, H. E., 1906 Brisbane St., Silver Spring, Md.
Peterson, H. L., 2715—31 St., S.E., Apt. 580, Washington 20, D. C.
Pifer, M. J., Sylvania Electric Products, Inc., 254 Rano St., Buffalo 7, N. Y.
Rideout, V. C., Electrical Engineering Department, University of Wisconsin, Madison 6, Wis.
Rosenthal, A. L., 235 Hillhurst Blvd., Toronto 12, Ont., Canada
Schottland, F. D., 76 Parkwold Dr., Valley Stream, L. I., N. Y.
Seifert, W. W., 623 Main St., Woburn, Mass.
Sherman, S., 113 Greenbelt La., Levittown, L. I., N. Y.
Smith, A. E., Box 234, McLean, Va.
Smith, C. V. L., Box 125, Lovettsville, Va.
Stavis, G., 104 Croton Ave., Ossining, N. Y.
Stein, R. S., 447 Bronxville Rd., Bronxville 8, N. Y.
Walker, H. E., 3226-J—34 St., Sandia Base, Albuquerque, N. Mex.
Whiffen, R. E., Ellendale Dr., R.F.D. 6, Towson 4, Md.

Admission to Senior Member

Ball, N. T., c/o Mark Swanson, Mason Hall, Alexandria, Va.
Covely, F. D., III, 600 Graisbury Ave., Haddonfield, N. J.
Davis, L. B., 1624 Alderson Ct., Owensboro, Ky.
George W. C., 2908 George Ave., Parma 29, Ohio
Gough, J. H., 6105 Lombard St., Cheverly, Md.
Herrick, M. P., Chief Engineer, Radio-Television Division, Stromberg-Carlson Co., Rochester 3, N. Y.
Hill, W. C., 4224 Fourth St., S.E., Washington, D. C.
Mason, V. V., Canadian Westinghouse Co., Ltd., Electronics Division, Longwood Rd., Hamilton, Ont., Canada

(Continued on page 90A)

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Seven precision cathode-ray oscilloscopes are available from Tektronix on an average shipping schedule of less than thirty days. These oscilloscopes are manufactured to the highest standards of quality and embody the very latest in practical circuitry. Each has characteristics desirable (in many cases, necessary) for a particular class of investigation or range of work. If from reading these brief descriptions you find you would like to know more, please write for the new Tektronix "long-form" catalog.

Type 315-D

Portable 3", wt. 36 lbs., 24 calibrated time bases 0.1 μsec/div to 5 sec/div (¼" divisions), 12 calibrated sensitivity positions 0.01 v/div to 50 v/div, vertical amplifier response dc to 5 mc, risetime less than 0.07 μsec (max. dc sensitivity 0.1 v/div). Trigger amplitude discriminator, direct-coupled unblanking, 5x sweep magnifier. 0.25 μsec signal delay. Square wave voltage calibrator. Flat-faced crt with 1.8 kv accelerating potential.

For 50-60 cycle supply—$770
For 50-800 cycle supply—$785

Type 511A-D

Continuously variable time base 0.1 μsec/cm to 0.01 sec/cm, 5x sweep magnifier. Vertical amplifier risetime 0.04 μsec, approximate bandwidth 5 cps to 10 mc, maximum sensitivity 0.125 v/cm, voltage calibrator zero to 100 v in 6 ranges. 0.25 μsec signal delay. New flat-faced 5" crt, 3 kv accelerating potential. Weight 55 lbs. Price—$845.

Type 512

Continuously variable time base 3 μsec/cm to 0.3 sec/cm, 1.0 sec/cm and 3.0 sec/cm available on special order, 5x sweep magnifier. Direct-coupled balanced vertical amplifier with differential input, risetime 0.2 μsec, sensitivity 0.005 v/cm to 50 v/cm at dc to 1 mc bandwidth—0.15 v/cm at dc to 2 mc, square wave voltage calibrator zero to 50 v in 9 ranges. 5" crt, 3 kv accelerating potential. Weight 56 lbs.

Price—$950.

Type 513

12 kv accelerating potential on 5" metallized crt. Continuously variable time base 0.1 μsec/cm to 0.01 sec/cm, 5x sweep magnifier. Distributed-type vertical amplifier, 0.025 μsec risetime, sensitivity 0.03 v/cm at 5 cps to 16 mc bandwidth—0.3 v/cm at dc to 18 mc. Square wave voltage calibrator, trigger rate generator, delayed gate and delayed trigger. 0.25 μsec signal delay. Self-contained, weight 79 lbs.

Price—$1650.

Type 514A-D

Continuously variable time base 0.1 μsec/cm to 0.01 sec/cm, 5x sweep magnifier, direct-coupled unblanking. Vertical amplifier risetime 0.04 μsec, sensitivity 0.03 v/cm at 2 cps to 10 mc bandwidth—0.3 v/cm at dc to 10 mc. Variable duty cycle square wave voltage calibrator zero to 50 v. 0.25 μsec signal delay. 6 cm undistorted vertical deflection. New 5" flat-faced crt, 3 kv accelerating potential. Weight 61 lbs.

Price—$950.

Type 517

24 kv accelerating potential on 5" metallized crt. Eleven calibrated time bases 0.01 μsec/cm to 20 μsec/cm, 2% accuracy. Distributed-type vertical amplifier risetime 0.007 μsec, sensitivity 0.1 v/cm. Panel switch to reduce accelerating potential to 12 kv giving twice the above sweep rates and sensitivity. Pulse-type voltage calibrator accurate within 4%, trigger rate generator. Approx. 0.12 μsec signal delay. Indicator and power supply mounted on Scope-mobile for mobility. Cathode follower probe, step attenuator, viewing hood included.

Price—$3500.

Type 524-D

Television Scope, designed for maintenance of TV transmitting and studio equipment. Variable delayed sweeps at the frame rate, field shift switch, sync separator, line frequency sweep, line indicating video output, variable duty cycle square wave calibrator. Continuously variable time base 0.1 μsec/cm to 0.01 sec/cm, 3x and 10x sweep magnifier. Vertical amplifier risetime 0.04 μsec, sensitivity 0.015 v/cm at 2 cps to 10 mc bandwidth—0.15 v/cm at dc to 10 mc, 0.25 μsec signal delay, 6 cm undistorted vertical deflection. New 5" flat-faced crt with 4 kv accelerating potential. Weight 61 lbs.

Price—$1180.

All dc voltages electronically regulated. Time base accuracy within 5% of full scale and calibrator accuracy within 3% of full scale except where otherwise stated. Prices f.o.b. Portland, Oregon.

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(Continued from page 89A)

Mellen, E. R., 614 Frelinghuysen Ave., Newark 5, N. J.
Millman, G. H., 319 Highland Rd., Ithaca, N. Y.
Ross, A. T., 12 Shadylawn Dr., Madison, N. J.
Saltzberg, B., 5143 N. Nashville Ave., Chicago 31, Ill.
Sprague, R. C., Sprague Electric Co., North Adams, Mass.
Wilcomb, E. F., 5 Willets Dr., Syosset, L. I., N. Y.

Transfer to Member

Barritt, R. C., 1200 Susquehanna Ave., W. Pittston, Pa.
Beach, H. W., 304 W. Berry St., White Apts. 304-A, Fort Wayne 2, Ind.
Bristow, F. E., 5628 N. Mascher St., Philadelphia 20, Pa.
Brooks, C. E., 33 W. Montgomery Ave., Hatboro, Pa.
Buder, S. H., 5123 Garden Ave., Merchantville 8, N. J.
Davis, R. W., Jr., 15 East Hill Dr., Doylestown, Pa.
Dorratcague, P. E., 6334—45 Ave., S.W., Seattle 6, Wash.
Forstall, E. L., 1500 Flat Rock Rd., Penn Valley, Narberth, Pa.
Frazier, R. L., 1001 Sanders Dr., Webster Grove 19 Mo.
French, J. C., 5243 S. La Grange Rd., La Grange, Ill.
Gabriel, E. Z., 223 Walnut St., Montclair, N. J.
Goldberg, H. S., 647 E. 14 St., New York 9, N. Y.
Gould, S., 211—109 Ave., Elmont, L. I., N. Y.
Gregal, J., 79 West St., Apt. 1-F, Worcester 2, Mass.
Grinnell, A. L., 12625 Archwood St., North Hollywood, Calif.
Gutterman, A. Z., 27 Harrogate St., Lido Beach, L. I., N. Y.
Heath, G. F., 1101 Wedgewood Rd., Baltimore 29, Md.
Hesse, H. R., General Electric Advanced Electronics Center, Ithaca, N. Y.
Hill, W. M., 603 N. Redmond, Berbany, Okla.
Hird, F. S., 224 S. Fifth St., Minneapolis 2, Minn.
Holtum, A. G., Jr., 605 Brinley Ave., Bradley Beach, N. J.
Honig, W. M., 6801 Bay Parkway, Brooklyn 4, N. Y.
Jagerman, D. L., 25-90—35 St., Astoria, L. I., N. Y.
Karns, K. B., Imperial, Nebr.
Lago, G. V., Department of Electrical Engineering, University of Missouri, Columbia, Mo.
Ling, O. T., 100 Henry St., Binghamton, N. Y.
Merrill, B., 34 East 76 St., New York 21, N. Y.
Molony, D. A., 482 Lincoln Ave., Highland Park, N. J.
Neuron, E., Box 504, R.F.D. 1, Camarillo, Calif.
Peterson, W. J., Box 344, Emporium, Pa.
Rosenberg, J. M., 1134 Norwood Ave., Oakland 10, Calif.
Schwartzberg, H. L., Philco Corp., G & I Division 28-A, 22 and Lehigh, Philadelphia 32, Pa.
Shnitzler, R. K., 11 Bilodeau Rd., Dorchester 21, Mass.
Smith, S. L., Box 3343, Beaumont, Tex.
Thomas, L. P., 5 Eldridge Ave., West Collingswood 6, N. J.
Vogel, E. H., 6630 Mardel, St. Louis 9, Mo.
Welch, D. P., 859 Highland Ave., Kenmore, Buffalo 23, N. Y.
Zweifel, F. A., 4020 Edison St., San Mateo, Calif.

Admission to Member

Andress, P. M., 3823 Oak Rd., Philadelphia 29, Pa.
Best, R. L., 68 Albany St., Cambridge 39, Mass.
Bolt, R. H., Acoustics Laboratory, Massachusetts Institute of Technology, 77 Massachusetts Ave., Cambridge 39, Mass.

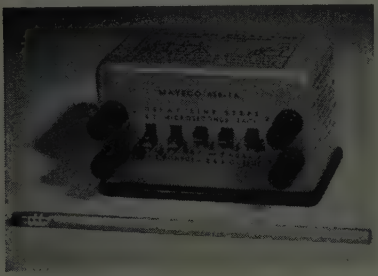
(Continued on page 92A)

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 85A)

Delay Lines

A laboratory-type delay line which is variable in additive increments of 0.02 microsecond each has been designed by **May Engineering Co.**, 6055 Lankershim Blvd., North Hollywood, Calif. It is used for electronic circuit work such as color television, instrumentation, pulse-forming networks, computer circuits, and so forth.

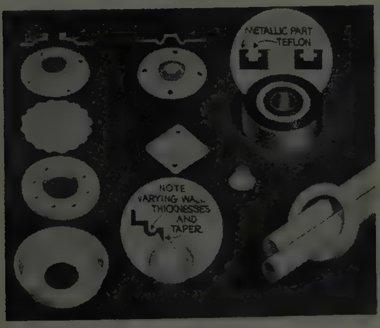


This instrument is the lumped-constant type with a total maximum delay of 1.0 microsecond measured at ½ amptitude with a rise time of 0.05 microsecond measured at 10 per cent and 90 per cent amplitude). The characteristic impedance is 50 ohms and maximum peak voltage is 500.

This delay line is constructed so that the individual switches each control a step of 20 coils and 20 condensers (1 per cent mica). Low attenuation is a feature of this unit. The overall dimensions are 2½×3×4½ inches. This delay line is in production and is priced at $75.00.

News Molding Techniques for Teflon

Recently perfected molding techniques have enabled the **Sparta Heat Treat Co., Plastic Div.**, East Sparta, Ohio, to mold DuPont's "Teflon" fluorocarbon in varying wall thicknessess, tapers, and undercuts formerly presumed impossible for that material.



Illustrated here are production molded diaphragm, spiral washers, discs, seals, and cup seals, representing fabrications in wall thicknesses of 0.047 inch to 0.031 inch and (upper right) a Teflon insert in brass.

Stock rods, tubing and sheeting are also produced by this firm specializing in Teflon.

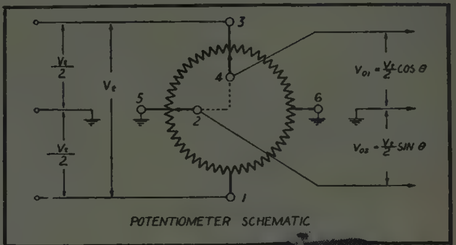
(Continued on page 96A)

TIC-TALKS
FEATURE

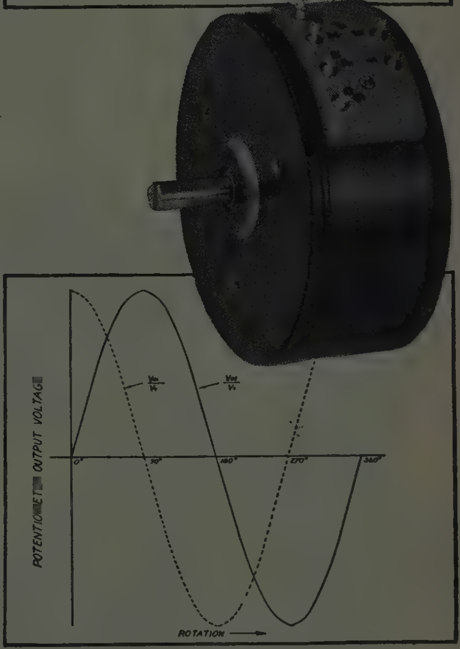
Something new in Precision Potentiometers . . .

. . . the standardization of a Non-Linear Precision Potentiometer, the *type* RVP3-S59 Sine-Cosine potentiometer, one of the many types standard with the Technology Instrument Corporation, performs two operations in a single potentiometer assembly . . . two wipers spaced 90 degrees apart yield both sine and cosine outputs.

1. Total resistance: 20,000 ohms plus or minus 5 per cent between terminals 1 and 3.
2. Accuracy: Plus or minus .5 per cent of the peak to peak amplitude.
3. Maximum voltage: Conservatively rated as 150 volts between terminal 1 and 3.
4. Life: Guaranteed for at least 500,000 complete cycles in either direction at 30 rpm.
5. Potentiometer base: Precision machined aluminum (originated by TIC) finished with corrosion resistant black Alumilite.
6. All fixed connections are soldered.
7. Wipers: Paliney spring wiper with double contact, for positive electrical connection, long wear and light torque.
8. Resistance Element: Karma wire with temperature coefficient of .00002 parts per degree centigrade.
9. Slip Rings: Inlaid coin silver slip rings. Paliney contacts on dual brushes for positive connection and low contact resistance.
10. Full humidity protection with type 76-S fungus resistant varnish.
11. Units may be ganged, using TIC's patented "Constrict-O-Grip" clamp rings which permit precise phasing with amazing ease.



POTENTIOMETER SCHEMATIC



TIC standard potentiometers have the same built-in precision and craftsmanship normally found only in custom-built products. Research, engineering and design facilities for special constructions and non-linear or linear functions are an integral part of TIC services. Submit your potentiometer problem, whether the need is for standard or custom design.

Engineering Representatives

Chicago, Ill. — UPtown 8-1141
Cleveland, Ohio — PRospect 1-6171
Waltham, Mass. — WAltham 5-6900
Boonton, N. J. — Boonton 8-3097
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Arnprior, Ont., Can. — Arnprior 400
Hollywood, Cal. — HOllywood 9-6305
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FLEXIBLE NYLON ADJUSTING SHAFT ELIMINATES BREAKAGE.
Picture-positioning lever. You specify mounting arrangement.

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Ft. Wayne 3, Indiana
Irv. M. Cochrane Co.
408 So. Alvarado St., Los Angeles, Calif.

MEMBERSHIP

(Continued from page 90A)

Bowker, R. F., Box 1250, Lanikai, Hawaii
Brubaker, J. W., 2328 Broadway, Summers Gyroscope Co., Santa Monica, Calif.
Byam, L. A., Jr., 60 Hudson St., New York 13, N.Y.
Cantrell, W. E., 813 Edney St., Fort Worth, Tex.
Castaldi, A. B., 1251 Belgrano, Buenos Aires, Argentina
Clemett, C. J., 103 Addison Rd., Caterham, Surrey, England
Croner, A. J., 119 W. Bataan Dr., Dayton 10, Ohio
Favors, H. A., 5154 Chimineas Ave., Tarzana, Calif.
Fisher, L. R., Emporium, Pa.
George, G. F., 9 Havana St., Roslindale, Mass.
Goldberg, A. A., 26 Stuyvesant Rd., Teaaneck, N. J.
Gurtz, V. J., 307 Brook Rd., Towson 4, Md.
Haefele, D. M., 9507 Meadowbrook Dr., Dallas, Tex.
Hammond, E. B., Jr., 13 Oak Ct., Merrick, L. I., N. Y.
Hunter, P. H., 875 Lockland Ave., Winston-Salem, N. C.
Knox, J. G., 157 Ann St., Clarendon Hills, Ill.
Lane, D. S., 522 Lowell Dr., Huntsville, Ala.
Locke, J. R., Jr., Roosevelt Ave., Liverpool, N. Y.
Maddox, R. G., 553 Berdan Pl., Long Beach, N. J.
Malmberg, P. R., Westinghouse Research Laboratories, E. Pittsburgh, Pa.
Mody, D. G., 7080 Neri Dr., La Mesa, Calif.
Moody, J. M., 3600 White Oak Rd., Cedar Rapids, Iowa
Mott, G. R., The Haloid Co., 282 Hollenbeck St., Rochester 3, N. Y.
Noll, P. W., 3 Wakefield St., Worcester 5, Mass.
Park, G. I., Philco Tech Rep Division, 22 St. and Lehigh Ave., Philadelphia 32, Pa.
Phelps, J. H., R.F.D. 3, Livermore Rd., Owensboro, Ky.
Popaduk, G. F., 5208 Castor Ave., Philadelphia, Pa.
Ramberg, E. C., RCA Laboratories, Princeton, N. J.
Raybin, M. W., 777 N. Park Ave., Apt. 2, Pomona, Calif.
Riester, H. A., Jr., 419 Alcott St., Philadelphia 20, Pa.
Rodgers, R. C., Machine Design, Penton Bldg., Cleveland 13, Ohio
Rother, D. E., 1635 Waverly Way, Baltimore 12, Md.
Sandsmark, P. I., 17 Fernwood Rd., Summit, N. J.
Sasseen, J. H., Humble Oil and Refining Co., Box 2180, Houston, Tex.
Schwartz, S., 22A-038, Massachusetts Institute of Technology, Cambridge, Mass.
Stevens, R., 155 N. Bridge, San Gabriel, Calif.
Stoddart, T. W. H., 11–17 Brentcliffe Rd., Toronto 17, Ont., Canada
Toperzer, R. S., Hughes Aircraft Co., Patent Department, Florence Ave. at Teale St., Culver City, Calif.
Tressa, F. J., 246 Division Ave., Massapequa, N. Y.
Utman, R. E., Aviation Supply Office, PURS-B, 700 Robbins Ave., Philadelphia 11, Pa.
Veraa, C. F., 906 Edgerly Rd., Glen Burnie, Md.
Verschuur, J. J., 351 Carlaw Ave., Toronto, Ont., Canada
Walter, G. M., 1602 Mussula Rd., Towson 4, Md.
Wood, G. R., Jr., 214 Floyd St., Belleville 9, N. J.
Worley, A. R., c/o 110 AC & W Sq, 151st Tac Con Gp, Camp Edwards, Mass.

The following elections to the Associate grade were approved to be effective as of July 1, 1953:

Aina, C. O., 80 Ashmore Rd., London W. 9, England
Andrews, W. J., 1006 Edgemoor Ct., Lancaster, Pa.
Ash, E. A., Electronics Research Laboratory, Stanford University, Stanford, Calif.
Barry, A., 6254-B Cartier, Montreal 35, Que., Canada

(Continued on page 94A)

ZOPHAR
---WAXES
---COMPOUNDS

Zophar Waxes, resins and compounds to impregnate, dip, seal, embed, or pot electronic and electrical equipment or components of all types; radio, television, etc. Cold flows from 100°F. to 285°F. Special waxes non-cracking at —76°F. Compounds meeting Government specifications plain or fungus resistant. Let us help you with your engineering problems.

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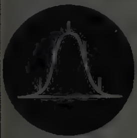
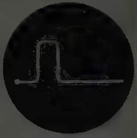
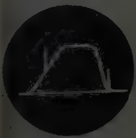
ZOPHAR MILLS, INC.
112-130 26th Street,
Brooklyn 32, N. Y.

NEED INSTRUMENTS TO SOLVE ELECTRONIC PROBLEMS?...


RADA-SWEEP
Radar IF Amplifier Alignment

MEGA-MATCH
Measurement of Reflection Coefficient


MEGA-PULSER
Transient Testing Video Amplifiers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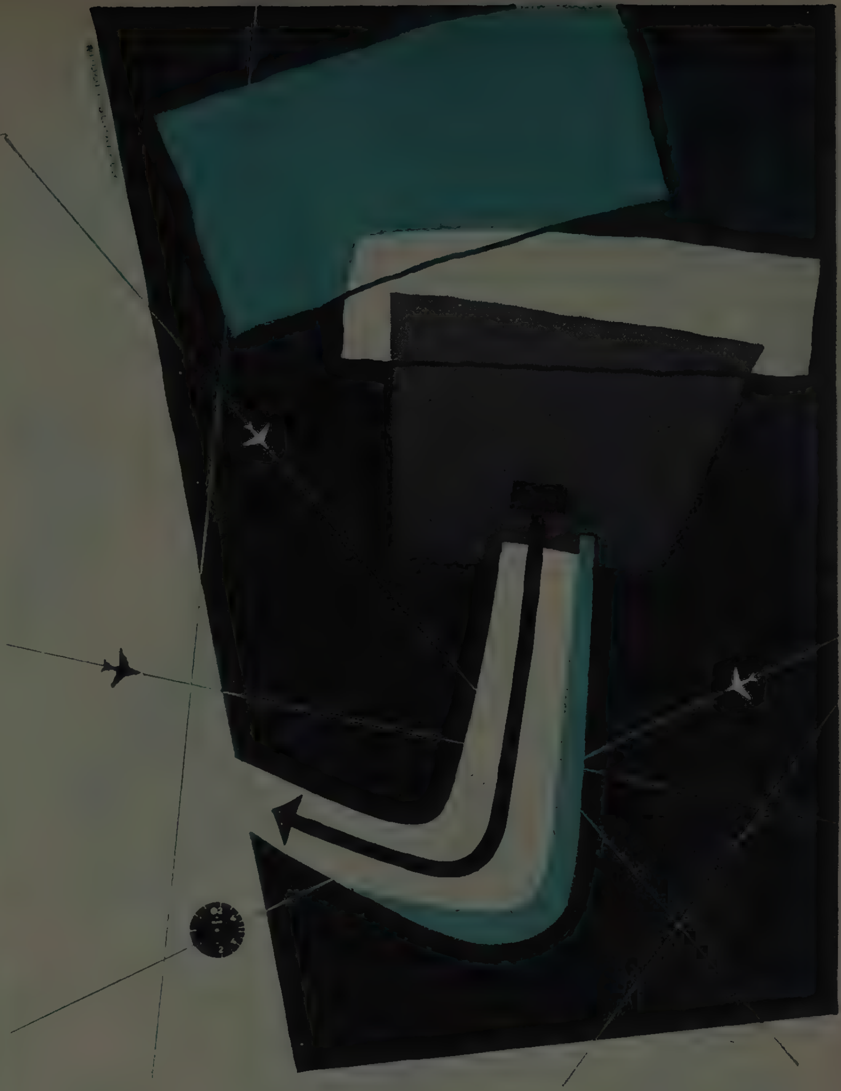

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Marafioti, J. J., Corning Glass Works, Corning, N. Y.

(Continued on page 95A)

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(Continued from page 94A)

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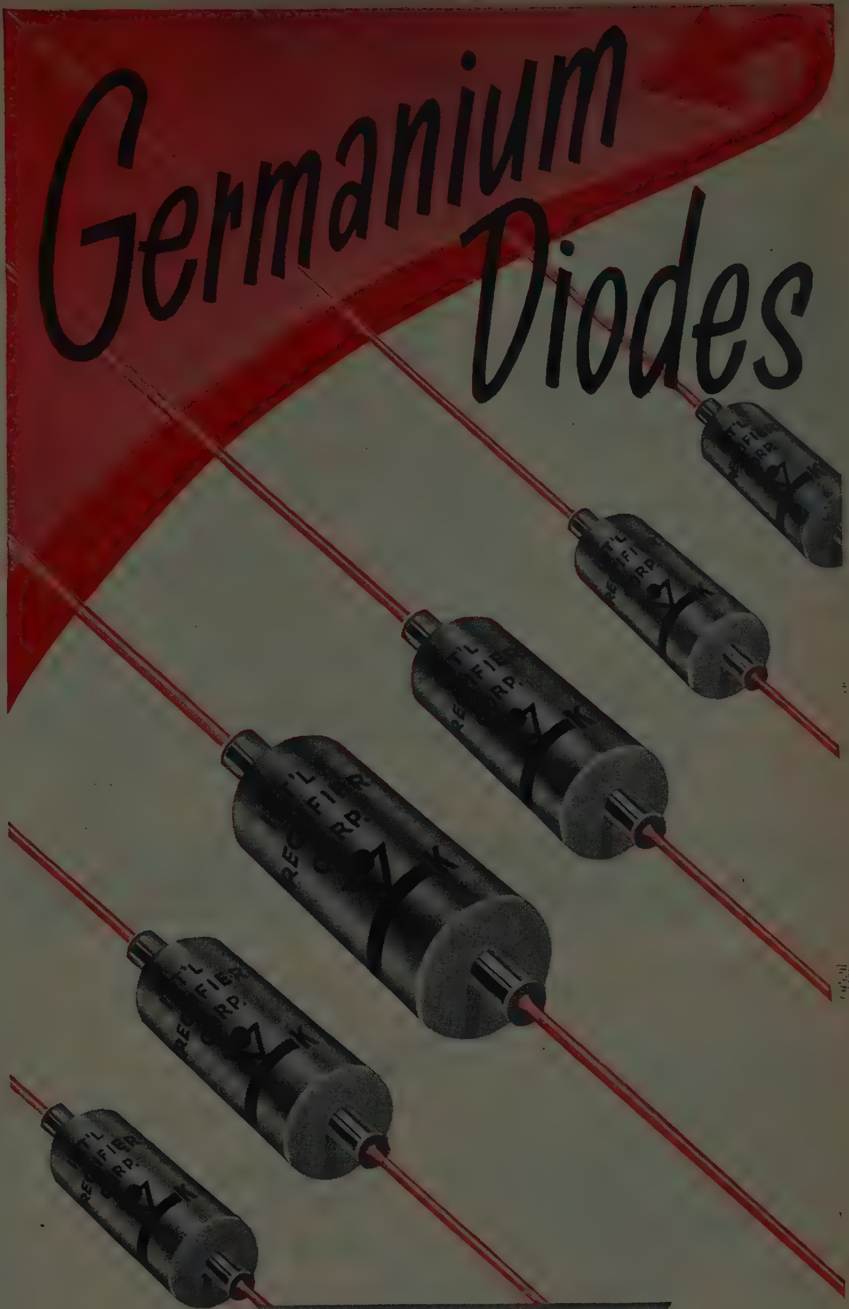
(Continued on page 96A)



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PHYSICAL STRUCTURE is ideally suited to impedance matching in standard coaxial line and waveguides.

FINISH. Coated with a special silicone varnish to protect the film.

SPECIFICATIONS
Resistance: 50 ohms standard, other values on request.
Tolerance: 5% or 10%
Wattage: 1/4 watt continuous duty at 25°C
Size: 1/16 inch diam. x 3/16 inch long
Terminals: Tinned sections 1/16 inch long
Film Length: Type R-063 — 1/16 inch
Type R-093 — 3/32 inch
Temperature Coefficient: approx. 0.0019 ohms/ohm/°C.
Power Sensitivity: Approx. 10 ohms/watt

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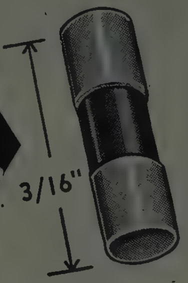

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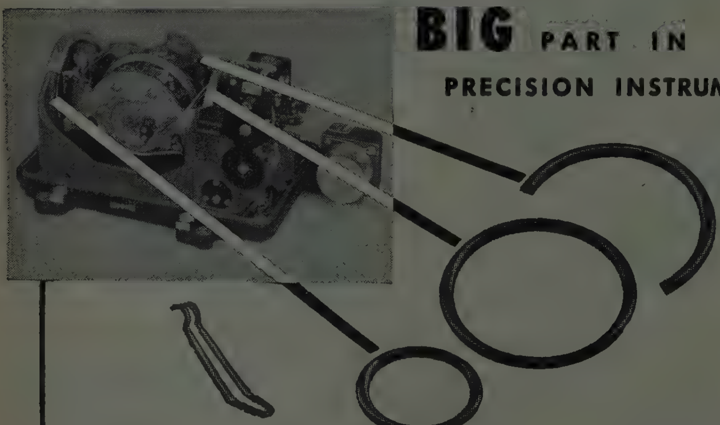

This Cageable Vertical Gyro, for use in stabilization and control systems of aircraft, guided missiles and radar scanners, manufactured by the Aeronautical Division of the Minneapolis-Honeywell Regulator Co., contains three doughnut potentiometers wound with NEY-ORO G high strength, precious metal resistance wire, contacted with wiper brushes of Paliney #7* (illustrated at far left).

Many other manufacturers of precision instruments specify Ney precious metal component parts for use as slip rings, brushes, wipers, commutator segments, etc. Ney Precious Metal Alloys have specific qualities which mean greater accuracy, longer life and resistance to most corrosive industrial atmospheres.

Call or write the Ney Engineering Department for assistance with your instrument problems.

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Watts, D. E., Patent Department, Remington Rand, Inc., South Norwalk, Conn.
Weaver, C. S., 1435—11 Ave., E., Twin Falls, Idaho
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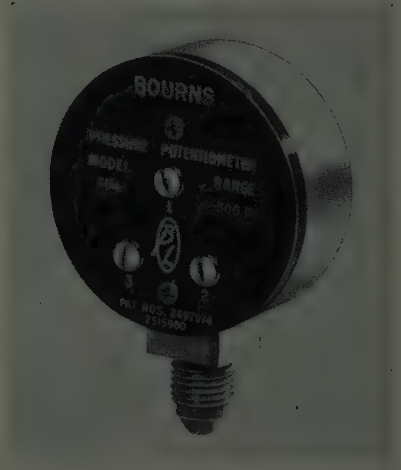
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 91A)

Miniature Gage Pressure Potentiometer

Bourns Labs, 6135 Magnolia Ave., Riverside, Calif., has developed new miniature-gage pressure potentiometers designed for remote measurement of hydraulic and pneumatic pressure in aircraft, guided missiles and other equipment.



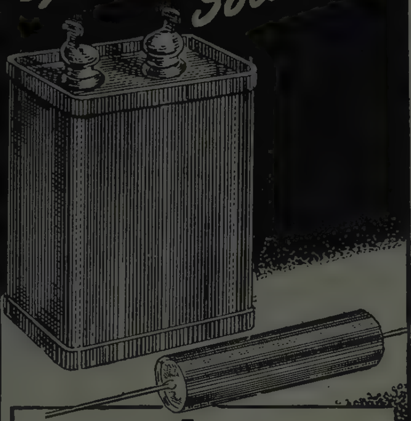
Actuated by the flexure of an accurately calibrated Bourdon tube, these precision instruments have a ½ watt electrical output which is transmitted to the wiper contact on the wire-wound resistance element by a ball-bearing supported linkage system. Available in ranges from 0–100 psi to 0–5,000 psi, they have a resolution of 0.4 per cent. A wide selection of standard resistance values add to their versatility.

The manufacturer claims that the instruments are not damaged by shock, vibration or nominal over-pressure.

(Continued on page 98A)

capacitors

Optimum Solution



unicon

Polystyrene Capacitors

Unicon Polystyrene Capacitors are characterized by extremely high insulation resistance and Q, and very low power factor and dielectric absorption.

Thus they are ideal elements for use in electronic analyzers and computers in tuned circuits and wave filters, as standards of capacitance, and in radiation detection and power supply circuits.

Unicon Polystyrene Capacitors are available in many different types of housings as well as uncased for potted assemblies. Standard sizes available for prompt delivery are listed below.

POLYSTYRENE LISTING

Cat. #		Cap.	WVDC	Style Housing	[illegible]
A1-1-	50	.05	100	1	1-1/2" Lx1/2" D
A1-1-	100	.10	100	1	1-1/2" Lx3/4" D
A3-1-	250	.25	100	3	1"x1-13/16"x1 H
A2-1-	500	.50	100	2	1-1/16"x1-13/16"x1-5/8" H
A2-1-	1000	1.00	100	2	1-1/16"x1-13/16"x2-3/4" H
A1-3-	1	.001	300	1	1-1/2" Lx1/2" D
A1-3-	2	.002	300	1	1-1/2" Lx1/2" D
A1-3-	5	.005	300	1	1-1/2" Lx1/2" D
A1-3-	10	.01	300	1	1-1/2" Lx1/2" D
A1-3-	20	02	300	1	1-1/2" L1/2" D
A1-3-	50	.05	300	1	1-1/2" Lx3/4" D
A1-3-	100	10	300	2	2-1/2" Lx3/4" D
A2-3-	250	.25	300	2	1-1/16"x1-13/16"x2-1/4" H
A2-3-	500	.50	300	2	1-1/16"x1-13/16"x3-1/4" H
A2-3-	1000	1.00	300	2	1-1/4"x2-1/2"x3-1/2" H
A2-3-	2000	2.00	300	2	1-1/4"x3-3/4"x3-3/4" H
A2-3-	5000	5.00	300	2	3-3/4"x4-9/16"x3-3/8" H
A2-3-	10000	10.00	300	2	3-3/4"x4-9/16"x5-1/8" H

NOTES:

A. Style 1 is hermetically sealed glass tube with metal end caps.
B Style 2 is hermetically sealed rectangular metal can. See JAN C-25 drawings of CP 70 cans for outline drawings.
C. Style 3 is hermetically sealed metal "bathtub" cans. See JAN -C-25 drawings of CP 53 cans for outline drawings.
D. 3% capacity tolerance is standard for polystyrene capacitors. Specials to ½% for Polystyrene capacitors are available at extra cost.

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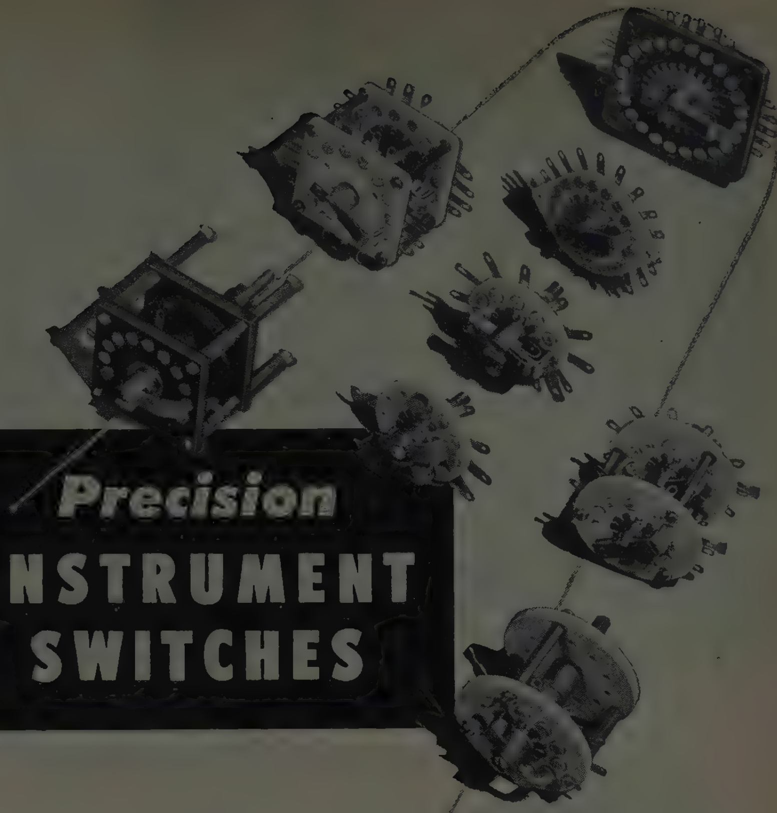

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There's nothing like stock models of Shallcross Instrument Switches for cutting equipment cost and design problems. For over 20 years Shallcross has made scores of basic switch types with countless variations for practically every electric-electronic application. Single or multi-deck types having up to 60 non-shorting positions are regularly produced. Contact resistance less than 0.001 ohm is a feature of many types. Several unique switches—usually "Special items" with other manufacturers—can quickly be supplied by Shallcross from stock. For a fast solution to any problem involving rotary switches, write to SHALLCROSS MANUFACTURING CO., 524 Pusey Ave., Collingdale, Pa.

Shallcross

North American F-86F Sabre Jet

VIBRATION DATA

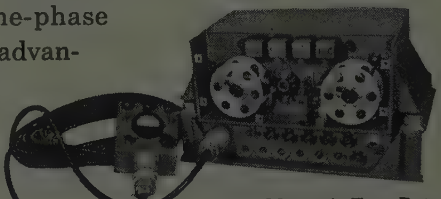
recorded by

14-channel Brush Magnetic Head

A precision Brush multichannel head assures faithful reproduction in the Davies magnetic tape data recorder, used to record vibration in jet aircraft and shock waves in seismic and oil field exploration.

This head has one synchronizing channel and 13 data channels, all of which record on a single tape 1¾″ wide. *Precision gap alignment* permits recording on one machine and playback on another machine with all signals in perfect time-phase relationship—an exclusive advantage of Brush multichannel heads.

For complete information, write Brush Electronics Company, Dept. F-8, 3405 Perkins Ave., Cleveland 14, Ohio.



Magnetic Tape Data Recorder, made by Davies Laboratories, Inc., Riverdale, Md., uses Brush Magnetic Heads for Accurate Recording.

BRUSH ELECTRONICS BRUSH ELECTRONICS **COMPANY**

INDUSTRIAL AND RESEARCH INSTRUMENTS
PIEZO-ELECTRIC MATERIALS • ACOUSTIC DEVICES
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Brush Electronics Company
is an operating unit of
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News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 96A)

High Speed Switching Commutator

G. M. Giannini & Co., Inc., 117 E. Colorado St., P.O. Box N, Pasadena 1, Calif., and East Orange, N. J., announces the availability of a new multi-channel, rotary, mechanical, high-speed switching commutator. The component is designed for use in radio and wire telemetering, multi-channel switching, and automatic rapid rate sampling. It is readily adaptable to give a wide variety of circuit patterns and number of channels.

G. M. GIANNINI & CO. INC.
EAST ORANGE N. J.

This commutator unit features dust-proof construction and noise-free operation throughout its operating life of at least 100 hours at 5 rps. A precision-ground, corrosion-resistant shaft mounted in ball bearings, insures close concentricity between shaft and mounting surface. With precious-metal contacts and brushes, and an external adjustment for phase relationship between poles, the commutator is designed to give clean electrical make-and-break, and either "break before make" or "make before break" types can be supplied.

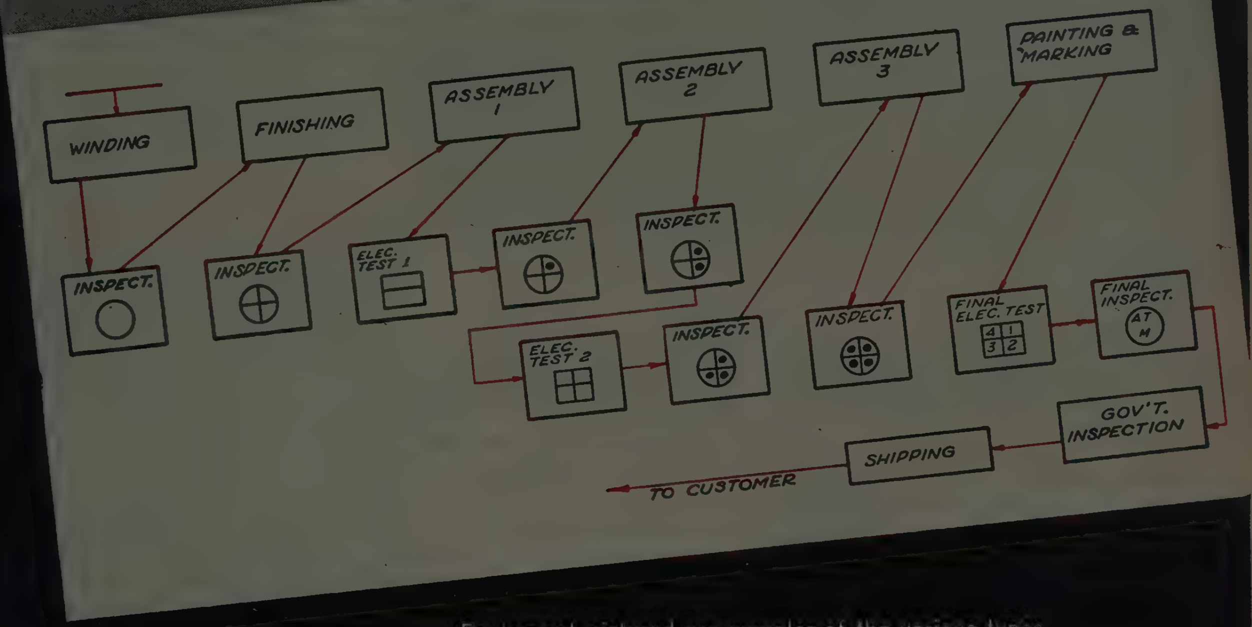
Typical specifications are: one or two poles with a maximum of 45 terminals on one pole and 15 terminals on the second pole. A maximum of 90 contacts on the one pole, and 30 on the other pole, can be had (two contacts per terminal). The off period is one degree minimum on one pole, three degrees minimum on the other pole.

The unit will meet performance specifications under 50G steady state acceleration and will operate under 10G's vibration up to 1,000 cps. Total weight is 3½ ounces.

(Continued on page 100A)

Quality First

Your Transformer Has To Pass These 10 Inspections


WINDING
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ASSEMBLY 2
ASSEMBLY 3
PAINTING & MARKING
INSPECT.
INSPECT.
ELEC. TEST 1
INSPECT.
INSPECT.
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Equipment pictured are samples of the various types of transformers manufactured by Aircraft Transformer Corporation, originators of the famed FORMFLEX

AIRCRAFT TRANSFORMER CORPORATION

MANUFACTURERS OF INDUCTIVE EQUIPMENT • LONG BRANCH, NEW JERSEY LONG BRANCH 6-6250

Plate and Filament

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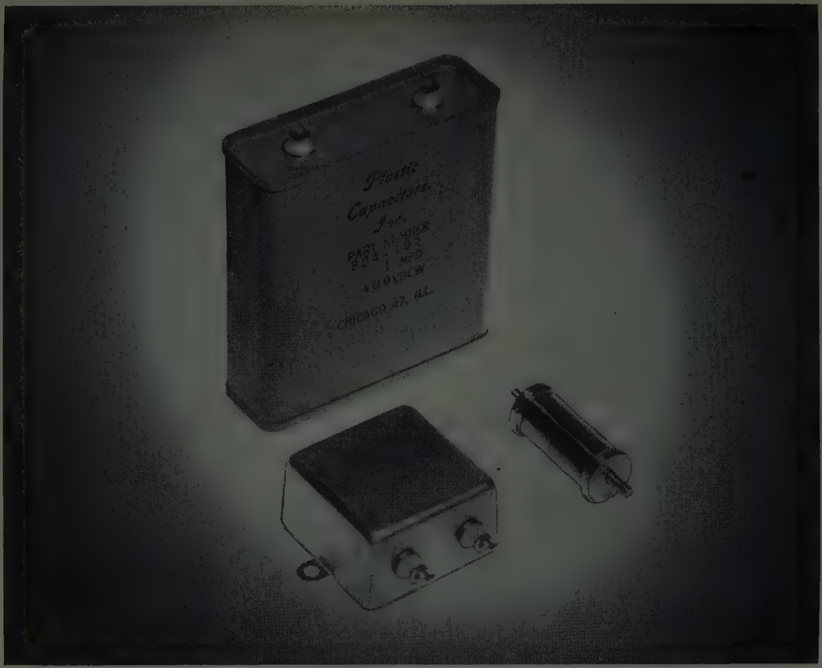
Representatives now located in principal cities throughout the country. Write for address of the one nearest you.



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Ultra capacitance stability with life
Very low dielectric absorption
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PC *Plastic Capacitors, Inc.*
2511 W. MOFFAT STREET, CHICAGO 47, ILLINOIS

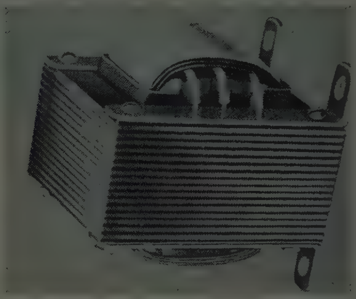
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 98A)

Filament Transformers

Ten new filament transformers for use with selenium rectifier applications have been introduced by the **Merit Coil and Transformer Corp.**, 4427 N. Clark St., Chicago 40, Ill.



These ten new transformers cover the complete range of present requirements for rectifiers used in industrial, laboratory, TV, and power applications.

Included in this group are units which correctly match the popular all-purpose Federal assembly kit.

These ten new Merit items, plus the nine units already in the Merit line, afford good coverage of selenium rectifier requirements.

Phase Meter

The Industrial Test Equipment Co., 55 E. 11 St., New York, N. Y., has introduced the Phazor Phase Meter Model 200A. This instrument permits phase shift measurements even though the input signal is complicated by noise and harmonic voltages. The device is useful in the measurement of phase shift in transformers, amplifiers, filters, and phase displacement networks. It can also be employed to measure either in-phase or quadrature voltage components. Under certain conditions phase shifts of the order of 0.01° can be measured with this instrument. Some of the features of this instrument are: self calibrating action, measures from 0 to 360°, readings not affected by noise and harmonics, high sensitivity, broad frequency range, high input impedance, wide input voltage range, electronic limiting to prevent overloading instrument.

Power input is 105–125 volts, 60 cps, 25 watts. Dimensions are: height 9 inches, width 15 inches, depth 8 inches.

(Continued on page 102A)

NEW!

GOLD BONDED

Germanium Diodes

Transistor Products germanium diodes, the smallest, *most efficient* diodes made, exhibit more uniform characteristics than any other diodes on the market. TP Germanium Diodes, with the Gold Bond feature, offer incomparable performance and dependability. The microminiaturized case, 3/16″ in diameter, gives new freedom of design and simplifies installation.

designed for Special TV Applications

X-16 for use as a frequency multiplier in TV UHF tuners.

X-18 for use as a second detector in TV receivers.

X-13 particularly useful as a clamper and balanced modulator in color TV circuits.

The above types are available NOW. Ready soon — a new cartridge-type UHF Mixer Diode.

Write Dept. E8 for descriptive literature

TRANSISTOR PRODUCTS, INC.
SNOW AND UNION STREETS, BOSTON 35, MASSACHUSETTS
AN OPERATING UNIT OF
CLEVITE CORPORATION

Cuts lines as fine as .008″
IT MAKES THE
IMPOSSIBLE JOBS
...easy!
It has been demonstrated that the "Airbrasive" Unit will do jobs that are impossible to accomplish by previously known methods. What is your problem?

THE S.S.White INDUSTRIAL "AIRBRASIVE" UNIT

THE S.S.WHITE Industrial "Airbrasive" Unit provides a fast, accurate and low-cost way to handle those tough, high precision jobs that are either impossible or impractical to do by conventional methods. For instance, it will:

- Cut germanium or other hard, brittle materials.
- Cut and shape fragile crystals.
- "Trim" surface coatings on printed circuits without harming the base material.
- Remove oxidation from circuits printed by the copper etch method.
- Produce fine matte finishes on metal and glass.
- Cut calibration marks on precision glassware.
- Do light deburring.
- Drill thin sections of glass and other material.
- Cut spiral bands on deposited carbon resistors.

The secret of the "Airbrasive" Unit's amazing precision and versatility is its unique principle of cutting by means of a high speed gas propelled stream of abrasive particles. The stream, which is directed at the work through a tiny orifice, produces a cool, shockless and controllable cutting effect which can be held to extremely close tolerances.

We'll be glad to conduct tests on samples of any material or parts you have and advise you as to the suitability of the "Airbrasive" cutting process for your needs. Write today. For full details... .

WRITE FOR BULLETIN 5307. *It provides a comprehensive outline of the many ways in which the Industrial "Airbrasive" Unit can be used. Send for a copy.*

THE S.S.White INDUSTRIAL DIVISION
DENTAL MFG. CO.
Dept. GA 10 East 40th St.
NEW YORK 16, N. Y.

WESTERN DISTRICT OFFICE: Times Building, Long Beach, Calif.

JK PRODUCTS

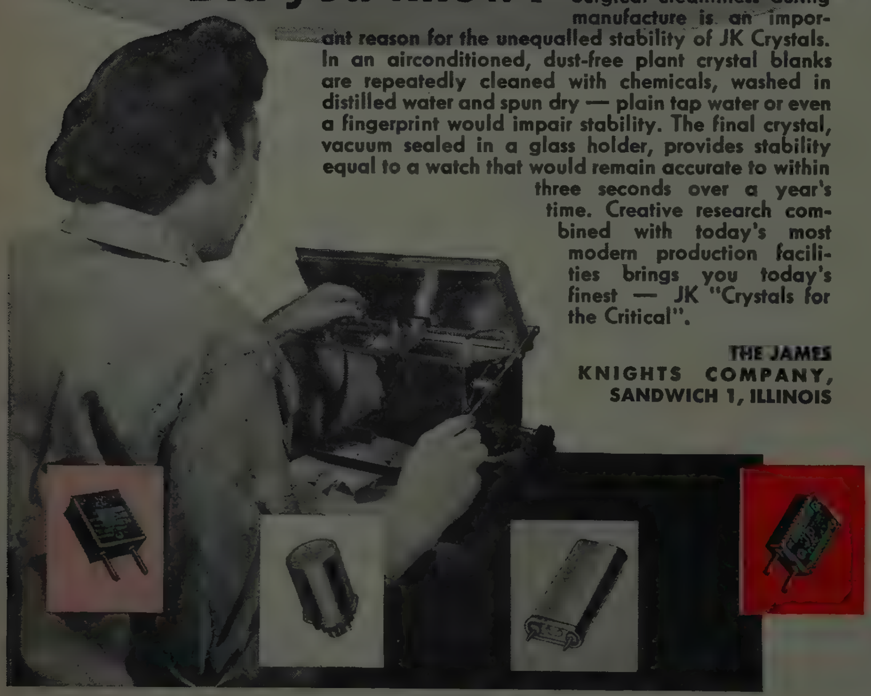
Speeding Electronic Progress through crystal research

The JK G-12 is a precision 100 kc G-T cut crystal intended for operation in Meacham Bridge and similar oscillators. Available for operation at series resonance or into large load capacities. Resistance approximately that of usual lamp used for amplitude stabilization, simplifying bridge circuit design. The JK G-12 is vacuum sealed. Equipped with octal base it is more convenient than usual "soldered-in" type of precision standard crystal. Suitable for transistor oscillators. Will fit JK 07EH temperature control unit. Consult us on specific applications.

JK STABILIZED G-12 CRYSTAL

Did you know? Surgical cleanliness during manufacture is an important reason for the unequalled stability of JK Crystals. In an airconditioned, dust-free plant crystal blanks are repeatedly cleaned with chemicals, washed in distilled water and spun dry — plain tap water or even a fingerprint would impair stability. The final crystal, vacuum sealed in a glass holder, provides stability equal to a watch that would remain accurate to within three seconds over a year's time. Creative research combined with today's most modern production facilities brings you today's finest — JK "Crystals for the Critical".

THE JAMES KNIGHTS COMPANY,
SANDWICH 1, ILLINOIS



News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 100A)

WESCON Technical Program Takes Form

General outlines of the technical papers program for the forthcoming WESCON (Western Electric Show and Convention) to be held in San Francisco, August 19–21, are crystallized into 17 sessions including approximately 85 papers, according to an announcement made by Bernard M. Oliver, Hewlett-Packard Co., WESCON papers chairman. WESCON is jointly sponsored by the IRE (7th region) and WCEMA (West Coast Electronic Manufacturers Association).

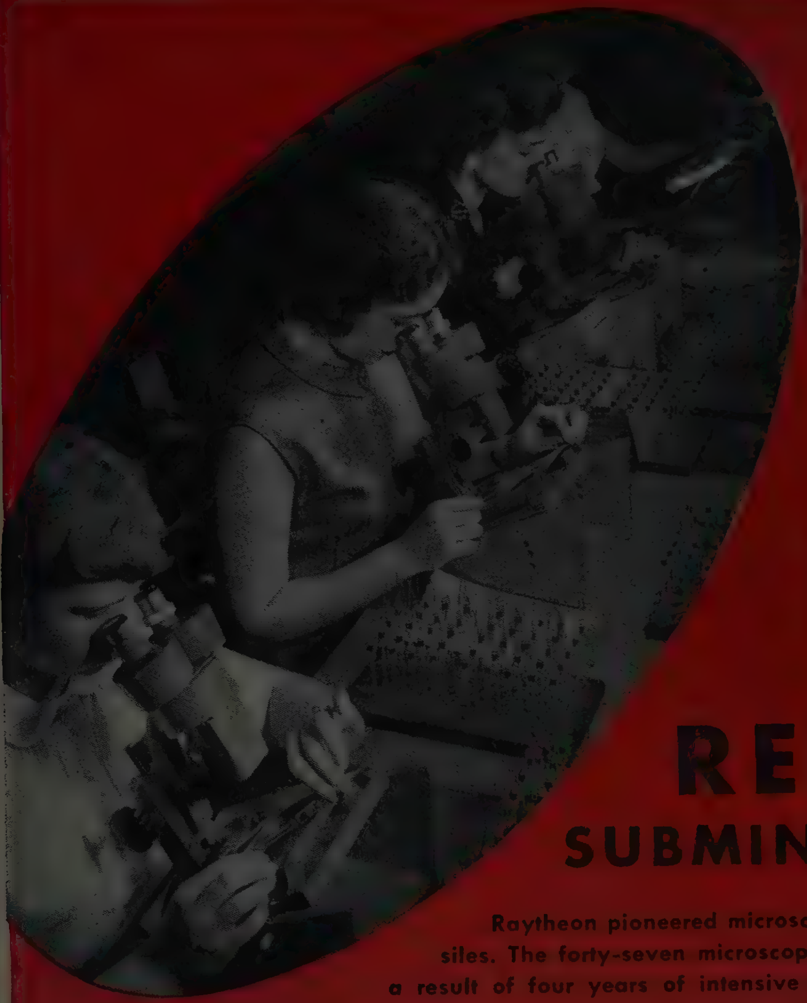
Still subject to minor changes, the program will include two sessions each on antennas, propagation, electron devices, circuits, and computers; and one session each on airborne electronics, microwave techniques, servos and telemetering, instrumentation, transistors, and nuclear radiation measurements. Of the evening sessions scheduled, one will be on medical electronics, another on audio, and a third on generalized topics of broad scope.

Additional technical interest will be imparted to the WESCON event by meetings on related subjects which are being scheduled to be held in the San Francisco Bay area preceding the WESCON program. One of these is a joint UCLA (University of California at Los Angeles) ONR (Office of Naval Research) information theory symposium in Berkeley August 17 and 18. Another is a special discussion of magnetic recording apparatus and techniques given for the Federal Court Reporters and sponsored by the San Francisco Section of the Audio Engineering Society.

Presiding chairman for the various sessions will be chosen by, and in many cases, from the following roster of subcommittee-men covering the various topics.

Antennas, Dr. Jack T. Bolljahn, Stanford Research Institute; Propagation, Allen M. Peterson, radio propagation laboratory, Stanford University; Airborne electronics, Allen Ellis, Stanford Research Institute; Circuits, Don Pedersen, electronic research laboratory, Stanford University, George Matthaei of the University of California, and Byron J. Bennett, Stanford Research Institute; Computers and Data Handling, Dr. Torben Meisling, University of California; Electron Devices, Marvin Chodorow, microwave laboratories, Stanford University, and Ted Moreno, Varian Associates; Instrumentation, Bruce Wholey, Hewlett-Packard Co.; Microwave Techniques, John R. Whinnery, University of California; Medical Electronics, Al Morris, Stanford University; Nuclear Radiation Measurement, Herbert Bright of U. S., Naval Radiological Laboratories, San Francisco; Servos and Telemetering, Otto J. N. Smith, University of California; Transistors, Howard Zeidler, Stanford Research Institute.

(Continued on page 104A)



47 MICROSCOPIC TESTS on all RAYTHEON RELIABLE SUBMINIATURE TUBES

Raytheon pioneered microscopic inspection of reliable tubes for missiles. The forty-seven microscopic tests were developed by Raytheon as a result of four years of intensive experience with visual inspection. This rigid multiple test procedure has been adopted as standard practice on every Raytheon Reliable Subminiature Tube. It assures utmost dependability of performance.

CK 5702 WA
CK 5703 WA

All these Raytheon Reliable Subminiature Tubes must pass microscopic inspection

Type	Description	Heater		Plate		Grid	Screen		Amp.	Mut.
		Volts	Ma	Volts	Ma	Volts	Volts	Ma	Factor	Cond.
CK5702WA	RF Amplifier Pentode	6.3	200	120	7.5	R_k = 200 ohms	120	2.5	—	5000
CK5703WA	High Frequency Triode	6.3	200	120	9.4	R_k = 220 ohms	—	—	25.5	5000
CK5744WA	High Mu Triode	6.3	200	250	4.2	R_k = 500 ohms	—	—	70	4000
CK5783WA	Voltage Reference		Operating voltage approximately 86 volts between 1.5 and 3.5 ma.							
CK5784WA	RF Mixer Pentode	6.3	200	120	5.2	−2	120	3.5	—	3200
CK5787WA	Voltage Regulator		Operating voltage approximately 100 volts between 1 and 25 ma.							
CK5829WA	Dual Diode	6.3	150		Max. Peak Inverse 360 volts. I_o = 5.5 ma. per plate					
CK6021	Medium Mu Dual Triode	6.3	300	100	6.5	R_k = 150 ohms	—	—	35	5400
CK6111	Medium Mu Dual Triode	6.3	300	100	8.5	R_k = 220 ohms	—	—	20	5000
CK6112	High Mu Dual Triode	6.3	300	100	0.8	R_k = 1500 ohms	—	—	70	1800
CK6152	Low Mu Triode	6.3	200	200	12.5	R_k = 680 ohms	—	—	15.8	4000

Note: All dual section tube ratings (except heater) are for each section.

Write for New Edition of
Raytheon Reliable Subminiature Tube Booklet

RAYTHEON®
Excellence in Electronics

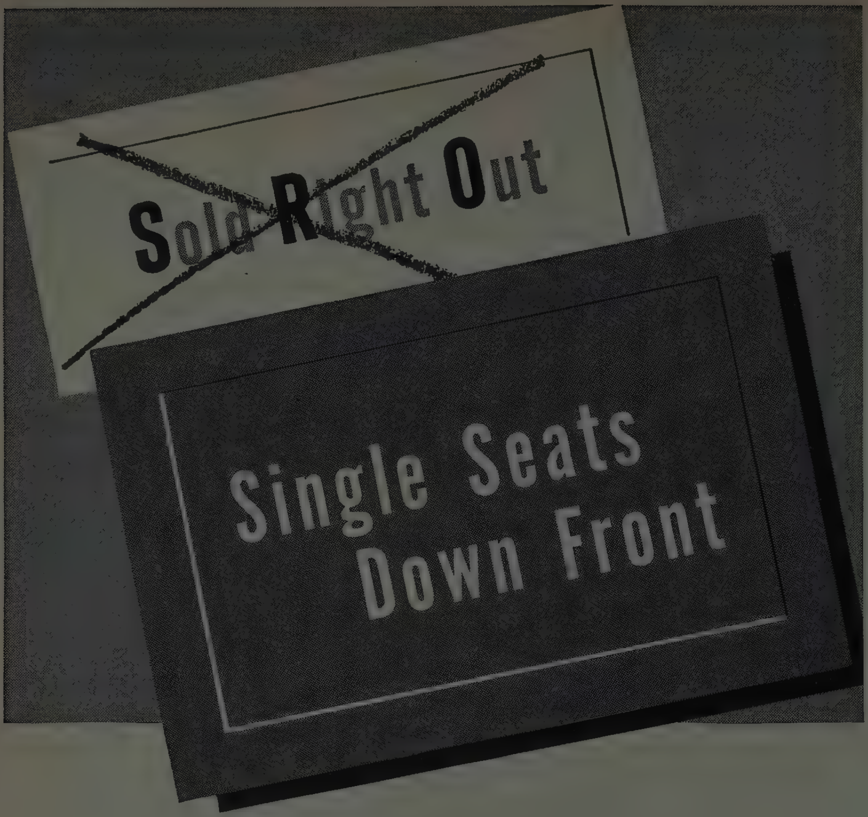
RAYTHEON MANUFACTURING COMPANY

Receiving Tube Division — for application information call

Newton, Mass. Bigelow 4-7500 ● Chicago, Ill. NAtional 2-2770 ● New York, N. Y. WHitehall 3-4980 ● Los Angeles, Calif. Richmond 7-5524

RAYTHEON MAKES ALL THESE:

RELIABLE SUBMINIATURE AND MINIATURE TUBES · GERMANIUM DIODES AND TRANSISTORS · NUCLEONIC TUBES · MICROWAVE TUBES · RECEIVING AND PICTURE TUBES



Our previous series of advertisements in this publication explained, in theatrical parlance, that our design and production facilities were pretty well "sold out" by the requirements of our present customers.

Now, we are happy to say (because we enjoy making new friends) that some of the heat has been taken off, and we are able to announce "Limited seating available"—as they say at the box office.

We shall be happy to talk with you about your present and/or future needs.

TRADE MARK

L. H. Terpening Company
DESIGN • RESEARCH • PRODUCTION
Microwave Transmission Lines and Associated Components
16 W. 61st St. • New York 23, N. Y. • Circle 6-4760

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 102A)

Subminiature Receiving Tubes

The General Electric Co., Tube Dept., Schenectady 5, N. Y., has announced the availability of two subminiature receiving tubes suitable for military application.

The GL-6111 is a medium-mu twin triode designed for general purpose amplifier applications. Each section has an individual cathode and is electrically independent. The tube also may be used as a combined oscillator and mixer in high-frequency circuits.

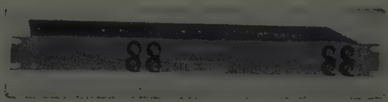
Characteristics and typical operation conditions for each section of the GL-6111 in use as a class Al amplifier are: plate voltage, 100 volts; cathode bias resistor, 220 ohms; amplification factor, 20; plate resistance, 4000 ohms, transconductance, 500 micromhos.

The GL-6112 is a high-mu twin triode suitable for use as an audio-frequency voltage amplifier or phase inverter. As a result of its low microphonic characteristics the tube can be employed at relatively low signal levels. The tube has a maximum noise output voltage of 0.025 volt with 100 volts on the plate.

Characteristics and typical operation conditions for each section as a class Al amplifier with 100 volts on the plate are: cathode bias resistor, 1500 ohms; amplification factor, 70; plate resistance, 39,000 ohms; transconductance, 1800 micromhos, plate current, 0.8 ma. With 150 volts on the plate: cathode bias resistor, 820 ohms; transconductance, 2500 micromhos; plate current, 1.75 ma.

Transformer Panels

The SKL Series 303 Transformer Panels to be used with the SKL Series 300 Variable Electronic Filters when a 600 ohm impedance is required have been designed by **Spencer-Kennedy Laboratories, Inc.,** 186 Massachusetts Ave., Cambridge 39, Mass.

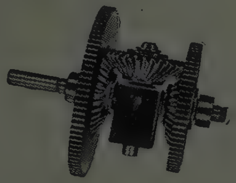


The Series 300 Transformer Panel comes in three models: Model 303A provides one 600 ohm output; Model 303B provides two 600 ohm outputs; and Model 303C provides two 600 ohm inputs and two 600 ohm outputs.

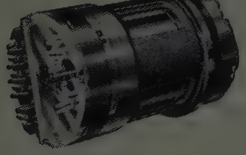
(Continued on page 106A)

FORD INSTRUMENT COMPONENTS

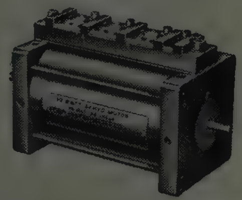
Ford Instrument Company makes the finest precision instruments and mechanisms for industry and the armed forces.



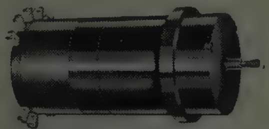
FORD MECHANICAL DIFFERENTIALS with single spider gear—available in 3/16," 1/4" and 5/16" shaft diameters—high accuracy, low friction.



FORD TELESYN UNITS are available in wide selection of sizes—proved precision accuracy, corrosion and fungus resistance mean better synchros when you specify Telesyn.



FORD SERVO MOTORS (60 and 400 cycles) for extremely low inertia and high frequency response. 1/5; ½; 1½; 2½; 5; 10-watt models in both low and high voltage, including magnetic amplifier controlled types.



FORD ELECTRICAL RESOLVERS available in sizes 23 and 31—interchangeability, temperature compensation −60°F to +160°F, highest accuracy, adaptability, 400 cycle frequency.

Ford Instrument also makes magnetic amplifier systems, computers, converters, mechanical integrators and other instruments and equipment. Write for more details.

11

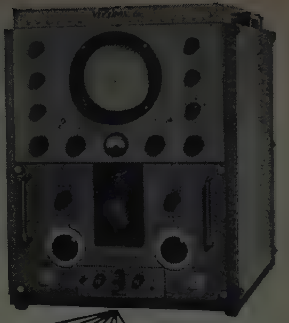
FORD INSTRUMENT COMPANY

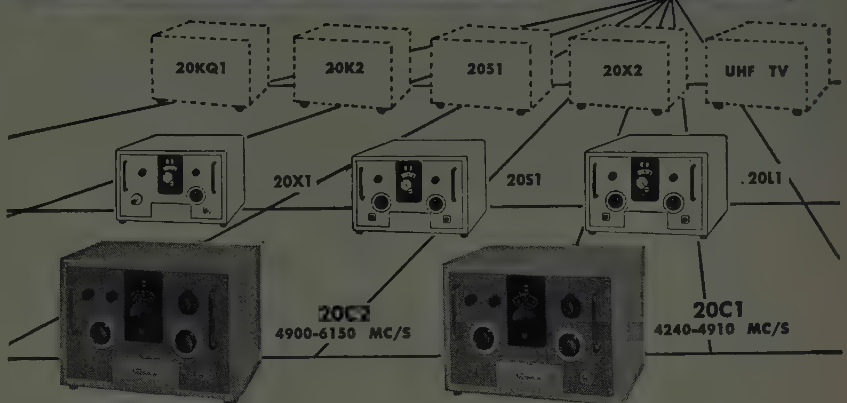
Division of The Sperry Corporation
31-10 Thomson Ave., Long Island City 1, N. Y.

TWO NEW C-BAND HEADS

for VECTRON'S

MICROWAVE SPECTRUM ANALYZER





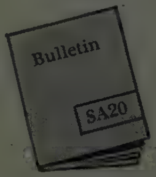
VECTRON'S two new R. F. Heads, 20C1 and 20C2, provide continuous coverage of microwave frequencies in C-band from 4,240 mc/s to 6,150 mc/s. They are engineered for immediate operation in Vectron's Spectrum Analyzer Chassis SA10 or SA20 . . . no conversion, no adaptation.

Specific Band Coverage . . . Vectron's R. F. Heads (L-band, S-band, C-band and X-band) are interchangeable in any Vectron SA20 Spectrum Analyzer. Display Unit and Heads can be purchased separately, as needed, to provide a wide choice of operating frequencies without the bulk and unnecessary expense of equipment which covers large areas in unused bands.

Early Delivery . . . Individual R. F. Heads and SA20 Analyzers are available for early delivery. Other new Heads are well along in development and will be announced soon. Send for Bulletin SA20 (see below) and specify the frequencies you need.

For Microwave Radar and Communications Equipment The Vectron SA20 Spectrum Analyzer presents visually the frequency distribution spectrum of the power output of pulsed or CW microwave oscillators and can be used as a sensitive R. F. detector for checks and measurements in the design, production and maintenance of microwave radar and communications equipment and components.

Vectron's development program includes additional R. F. Heads to cover microwave frequencies newly opened for military and civilian use. For information on these additional R. F. Heads and for complete engineering data, send for Bulletin SA20. Write today and be sure to specify the operating frequencies you need.



FEATURES

Large, clear 5" oscilloscope pattern

Standard bezel to accept camera, hood or filter

Minimum number of controls . . . maximum operating convenience

Double conversion assures I. F. alignment stability

Built in regulated supply for Klystron oscillator

Easy access for maintenance or adjustment

New Provision for D.C. Filament supply for Klystron oscillator

SPECIFICATIONS

Overall Gain — 130 decibels

Sensitivity — At least 60 dbm for 1 usec. pulse — 80 dbm for CW

IF Bandwidth — Choice of 50 kc, recommended for CW and 0.2 to 2 usec. pulse widths, or 20 kc bandwith to 5 usec.

Sweep Frequency — 10 to 30 cps standard — available to 2 cps and with long persistence tube

Power supply 105-125V, 50-400 cycles

New Low-noise 20kc IF strip for higher useable gain.

VECTRON, inc. *Electronic and Electro-Mechanical Equipment*

402 MAIN STREET, WALTHAM 54, MASS.

Vectron invites your inquiries regarding the development and manufacture of:

Gyro-mechanisms
Gyro-stabilized Platforms
Gyros and Gyro Systems
Aircraft Instruments
Precision Mechanical Devices
Computers and Calculators
Gear Assemblies
Test Instruments
Synchros and Servomechanisms
Electronic Systems
Precision Electronic Components
Communication Networks and Filters
Microwave Equipment
Microwave Test Equipment
Radar Units and Systems
Echo Boxes

Computation

The Arab
who first wrote zero

And
the mathematic clue

To
nothing at all is a hero

Who
never got his due

Unless
he sought

To share
the eternal naught.

—from
"Stars and Atoms
HAVE NO SIZE"
by A. M. SULLIVAN
(by permission)

MAXSON

Maxson offers top-caliber engineers exceptional opportunities

MAXSON ENGINEERING specializes in the development and manufacture of computers for precise determination of complex problems. Our 18 years experience in navigation, ordnance, and electronics is implemented by 600 engineering personnel, 2500 other employees, and 400,000 feet of laboratory and plant space.

This adaptable organization is available to augment yours. For details, see the latest Report of Maxson Facilities.

The W. L. MAXSON CORPORATION
UNIMAX SWITCH DIVISION • LANGEVIN MANUFACTURING CORPORATION
460 WEST 34TH STREET NEW YORK 1, NEW YORK
Plants at Long Island City, N.Y. and Old Forge, Pennsylvania

New UHF SWEEP GENERATOR for UHF TV Production Testing

TYPE 1211

The Type 1211 UHF Sweep Generator has been specifically designed to rapidly and accurately align UHF Television heads, converters and complete receivers,

SPECIFICATIONS

FREQUENCY COVERAGE: 450 to 900 MC. Dial calibrated in 36 MC steps. **BANDWIDTH:** Constant bandwidth of 50 MC over entire spectrum. Can be adjusted to narrower bandwiths, with internal controls. **MARKERS:** Pulse type crystal controlled accurate to 0.02% spaced 36 MC throughout the 450 to 900 MC spectrum. **OUTPUT:** At least 1 volt across a 75 ohm load. **ATTENUATOR:** Electrostatically coupled piston type, range approximately 80 db. **AUXILIARY OUTPUT SIGNALS:** 1. Automatically phased sine sweep for X axis of scope. 2. Marker pulses either plus or minus polarity, continuously variable in amplitude.

PRICE $950.00 F.O.B. PLANT

Manufacturers of a Complete Line of TV Test Equipment
TIC Tel-Instrument Co. Inc.
50 PATERSON AVENUE • EAST RUTHERFORD, N. J.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 104A)

Two New TV Tubes

Production of two new cathode-ray tubes for television application, Type 24CP4 and Type 24CP4A, has been announced by **The Rauland Corp.,** 4245 N. Knox Ave., Chicago, Ill.

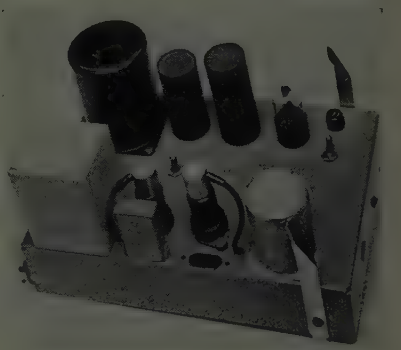
The new types are rectangular, glass, magnetic focus and magnetic deflection direct-view picture tubes. The gun is used with an external single-field ion trap magnet; and the tube has an external conductive coating that acts as a filter capacitor when grounded.

A gray, filter-glass faceplate in each is employed to improve contrast in the presence of ambient light. The aluminized tube, Type 24CP4A, counterpart of Type 24CP4 gives additional light output because of its metal-backed screen.

For more information or a specification sheet write to Rauland.

Three Audio Accessories

Three audio equipment items have been announced by the **Engineering Products Dept., RCA Victor Div., Radio Corporation of America,** Camden, N. J.



The new utility amplifier (Type BA 12A) is a two-stage high-fidelity unit whose high output signal level allows its use as a line amplifier, turntable booster, microphone preamplifier or isolation amplifier. The amplifier has frequency response of ±1 db 30 to 15,000 cps. Other features include low distortion and low hum level, self-contained power supply,

(Continued on page 108A)

DALIS DELIVERS

IMMEDIATELY... FROM STOCK...

The NEW

RAYTHEON

Junction-Type

TRANSISTORS

ORDERS FILLED IN SEQUENCE OF RECEIPT... SO,

Don't Delay!

PHONE: ALgonquin 5-3000

WRITE or WIRE: Dept. IND

TELETYPE: N. Y. 1-2482

Write for Free Technical Bulletins

Seal of HLD *Service*

H. L. DALIS INC.

WHOLESALE DISTRIBUTORS
ELECTRONIC EQUIPMENT

175 VARICK ST., NEW YORK 14

SERVING THE TRADE FOR OVER A QUARTER CENTURY

We've Stepped-up Production!

NOW Sangamo HUMIDITITE* Mica Capacitors are available in Quantity

When Sangamo HUMIDITITE Molded Mica Capacitors were first put on the market, the great interest shown in these remarkably moisture resistant capacitors far exceeded our expectations. We have increased our manufacturing facilities and our production capacity . . . initial demands have been met . . . and we can now handle quantity orders for Humiditite Micas with full assurance that delivery requirements will be met.

*what is HUMIDITITE?

Humiditite is the very effective new plastic molding compound, developed by Sangamo, that gives Sangamo Mica Capacitors moisture resistance properties far superior to any others on the market.

Sangamo Humiditite Micas, under the standard moisture resistance tests described in MIL-C-5A (proposed) Specification, tested in excess of 50,000 megohms—more than 500 times the specification requirements.

Humiditite is just another example of the advanced engineering that enables Sangamo to meet the existing and future needs of the electronic industry. For additional information about HUMIDITITE, write for Engineering Bulletin No. TS-111.

Those who know... SANGAMO choose Sangamo

SANGAMO ELECTRIC COMPANY

MARION, ILLINOIS

SO 53-4

STANDARD SIGNAL GENERATOR

MODEL 84—300-1000 Megacycles

OUTPUT VOLTAGE: Continuously variable from 0.1 to 100,000 microvolts. Output impedance, 50 ohms.

MODULATION: Sine Wave: 0-30%, 400, 1000 or 2500 cycles. Pulse: Frequency, 60 to 100,000 cycles. Width, 1 to 50 microseconds. Delay, 0 to 50 microseconds. Sync. output, up to 50 volts, either polarity.

POWER SUPPLY: 117 volts, 60 cycles. (Also available for 117 volts, 50 cycles; 220 volts, 60 cycles; 220 volts, 50 cycles.)

DIMENSIONS: 12" high x 26" wide x 10" deep, overall.

WEIGHT: Approximately 135 pounds, including external line voltage regulator.

Laboratory Standards

MANUFACTURERS OF

- Standard Signal Generators
- Pulse Generators
- FM Signal Generators
- Square Wave Generators
- Vacuum Tube Voltmeters
- UHF Radio Noise & Field Strength Meters
- L-C-R Bridges
- Megohm Meters
- Megacycle Meters
- Intermodulation Meters
- TV & FM Test Equipment

MEASUREMENTS CORPORATION

BOONTON · NEW JERSEY

AGASTAT

Compact . . . Dust-Proof

TIME DELAY RELAYS

solenoid actuated—pneumatically timed

Introduces time delays into a-c or d-c circuits. Easily adjusted to provide delays ranging from 0.1 second to five or more minutes.

The AGASTAT is small, light, and operates in any position. Dust-proof timing chamber assures long operating life with a minimum of maintenance.

Write for Bulletin.
Dept. A1-87

AGA

Division of Elastic Stop Nut Corporation of America
1027 Newark Avenue, Elizabeth 3, New Jersey

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 106A)

plug-in electrolytic capacitors, plug-in chassis, and compact size.

The new speaker mechanism (LC CA) and associated wall housing are designed specifically to work together in providing economical broadcast monitoring. The combination consists of a 10-inch permanent-magnetic type loudspeaker mechanism, and a sloped housing for 30 or 60° wall mounting.

The microphone boom and stand permits accurate studio microphone placement for best pickup. The boom telescopes from 6 feet 10 inches to 18 feet, while the stand is adjustable from 4 to 8 feet in height. It can be rotated 360° and perfect balance is maintained by an adjustable 25-pound steel counterweight which slides on the boom and locks securely at any position.

Literature fully describing the new audio equipment is now available from the Broadcast Equipment Section, RCA Victor, Camden 2, N. J.

Electronic Standard Cell

A new Electronic Standard Cell which furnishes a stable power supply from a 28 volt dc input is announced by the **Hastings Instrument Co., Inc.,** Super Highway at Pine Ave., Hampton, Va.

The cell, designed for aircraft use, has proven stable in repeated tests. When used intermittently, the cell has less than ±0.15 per cent change in output. The unit also shows less than 0.1 per cent total output change when used continuously. An ac ripple of less than 1 millivolt is another feature.

ELECTRONIC
STANDARD CELL

With such stability, the cell will meet the requirement of those applications requiring a stable output within the range of 0 to 100 volts and 0 to 30 milliamperes. Output voltage can be set anywhere within this range at the factory according to the users' specifications.

Because of its low output ripple, the unit may be used with all types of recording oscillographs or for instruments which measure temperature, power, motion, strain, velocity and mass flow.

(Continued on page 110A)

OPTICAL SYSTEMS

INDUSTRIAL PERISCOPES

KOLLMORGEN

DESIGN

DEVELOPMENT

MANUFACTURE

For nearly half a century Kollmorgen has designed, developed and manufactured precision optics and optical systems for industry and the military.

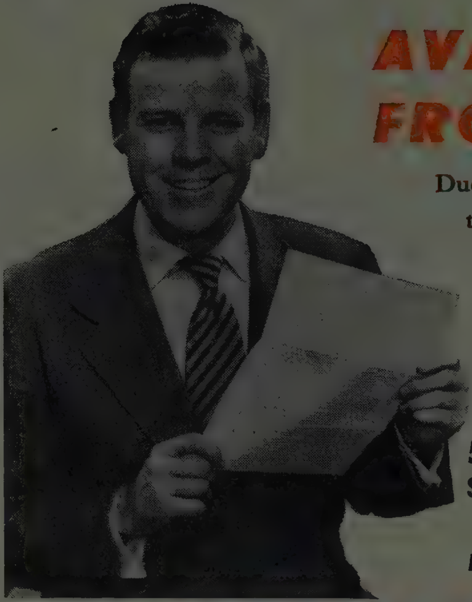
We have the engineering "know-how", the design personnel and the manufacturing capacity to help you solve your optical problem.

KOLLMORGEN *Optical* CORPORATION

Plant: 347 King Street • Northampton, Mass.

New York Office
30 Church Street
New York 7, N. Y.

NOW for the first time in *Over 10 Years* BOGUE 400 CYCLE GENERATORS



AVAILABLE FROM STOCK!

Due to the extremely high production at our newly enlarged plant, we are able to offer for immediate shipment from stock "World Famous" Bogue Hi-Cycle Generators in the following sizes:

5, 10, 20 and 50 KVA
Single Phase & Three Phase Output
220/440 Volt Input

Deliver 400 cycles regardless of load and input variations

LOW HARMONICS CLOSE VOLTAGE REGULATION

Our engineering department will be glad to supply full specifications on stock units as well as on special units to meet all Hi-Cycle requirements.

SINCE 1892 SPECIALISTS • IN FINE MOTORS & GENERATORS
Bogue

BOGUE ELECTRIC

The Authority on High Cycle Power

52 IOWA AVENUE • PATERSON 3, NEW JERSEY

New 800-2600 mcs FREQUENCY METERS
Lightweight — Portable Units

SPECIFICATIONS
1. ACCURACY
Better than .05% from 20°F to 120°F
2. SENSITIVITY
Usable indication with 1 milliwatt input
Adjustable for higher levels
3. INDICATOR
50 Microammeter
4. INPUT
50 Ohm Type N Connector
5. EXTERNAL DC OUTPUT
Pin Jacks
6. EXCURSION OF MICROMETER
One-half inch
7. MICROMETER SCALE
at 1000 Mc —
1 Division equals 290 KC
at 1400 Mc —
1 Division equals 350 KC
at 2000 Mc —
1 Division equals 450 KC
at 2600 Mc —
1 Division equals 555 KC
8. EXTERNAL SIZE
6½ x 9¾ x 7"
9. WEIGHT
Four pounds

...For Field and Laboratory Use!

Models
FS-C-171-A 800-1200 MCS.
FS-C-172-A 1200-1600 MCS.
FS-C-173-A 1600-2250 MCS.
FS-C-174-A 1700-2600 MCS.

Units consist of cavity body, micrometer control, crystal, suitable connectors and calibration chart. Write for specifications and prices.

(Cavity units are also available for custom housing)

frequency standards
P. O. Box 504.
Asbury Park, New Jersey

LFE

SOLID DELAY LINES STOP TIME . .

Suitable for video integration, computers, time markers, moving target indication, etc.

LFE Solid Delay Lines offer important advantages in obtaining precise delay intervals for pulse or modulated signals:

- Wide ranges of delay
- Low attenuation
- Low spurious response
- Wide bandwidth
- Smooth pass band
- Wide temperature range
- Minimum size and weight
- Rugged construction

For complete information, write: Specialties Division

LABORATORY for ELECTRONICS, INC.
75-4 PITTS STREET • BOSTON 14, MASS.

PRECISION ELECTRONIC EQUIPMENT · OSCILLOSCOPES · MAGNETOMETERS · COMPUTERS · MICROWAVE OSCILLATORS · MERCURY DELAY LINES

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

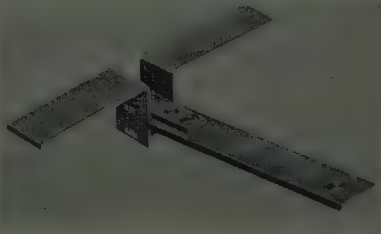
(Continued from page 108A)

Power Amplifier

A new direct-coupled amplifier, capable of supplying substantial output power, has been announced by the **Southwestern Industrial Electronics Co.,** 2831 Post Oak Rd., P.O. Box 13058, Houston 19, Texas. The Model B Power Amplifier has essentially flat response from 0 to 20,000 cps and will deliver 6.25 watts into a 100 ohm load when driven by a 10 volt signal. All types of distortion are held to extremely low levels. The Model B may be used to amplify the output of SIE or other oscillators and signal generators in order to drive Helmholtz coils, underwater sound transducers, servo systems, and other power-consuming devices. It may also be used to provide power from the output of straingage bridges, geophones, differential transformers, direct-coupled voltage amplifiers, photo-electric simulators, and various other transducers in order to facilitate recording and analysis. It is particularly useful in dealing with transient phenomena. The amplifier is mounted on two standard 8¾ by 19 inch rack panels, weighs 150 pounds, and operates on 115 v 60 cps ac at 450 watts.

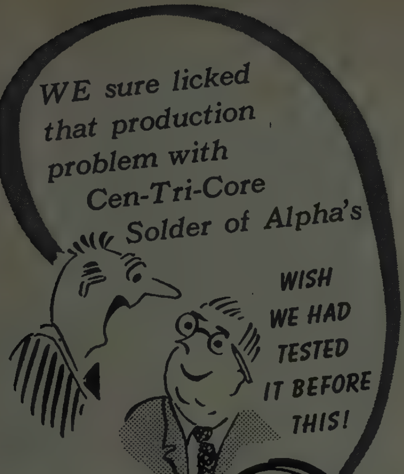
Anti-Pincushion Correcting Magnet

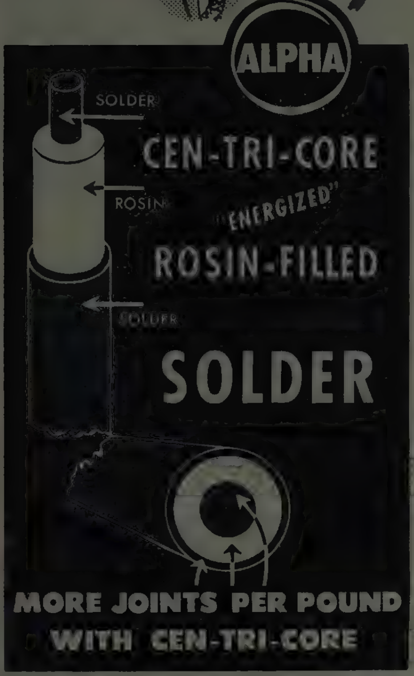
A new anti-pincushion Alnico correcting magnet is now available from **Heppner Manufacturing Co.,** Round Lake, Ill. The new magnet is designed to permanently correct pincushioning effect on 21-, 24-, and 27-inch TV picture tubes. Any type of mounting bracket specified by the customer can be furnished.



(Continued on page 143A)

ARE **YOU** IN THE **KNOW** TOO ?


WE sure licked that production problem with Cen-Tri-Core Solder of Alpha's
WISH WE HAD TESTED IT BEFORE THIS!


ALPHA
SOLDER
CEN-TRI-CORE
ROSIN
"ENERGIZED"
ROSIN-FILLED
SOLDER
SOLDER
MORE JOINTS PER POUND WITH CEN-TRI-CORE

Exclusive Features

- guarantees against rosin voids or skips
- eliminates cold joints and rejects
- available in eight core sizes
- solders to plated or oxidized parts
- simultaneous "wetting flow" and take
- surpasses federal specifications for non corrosiveness and purity

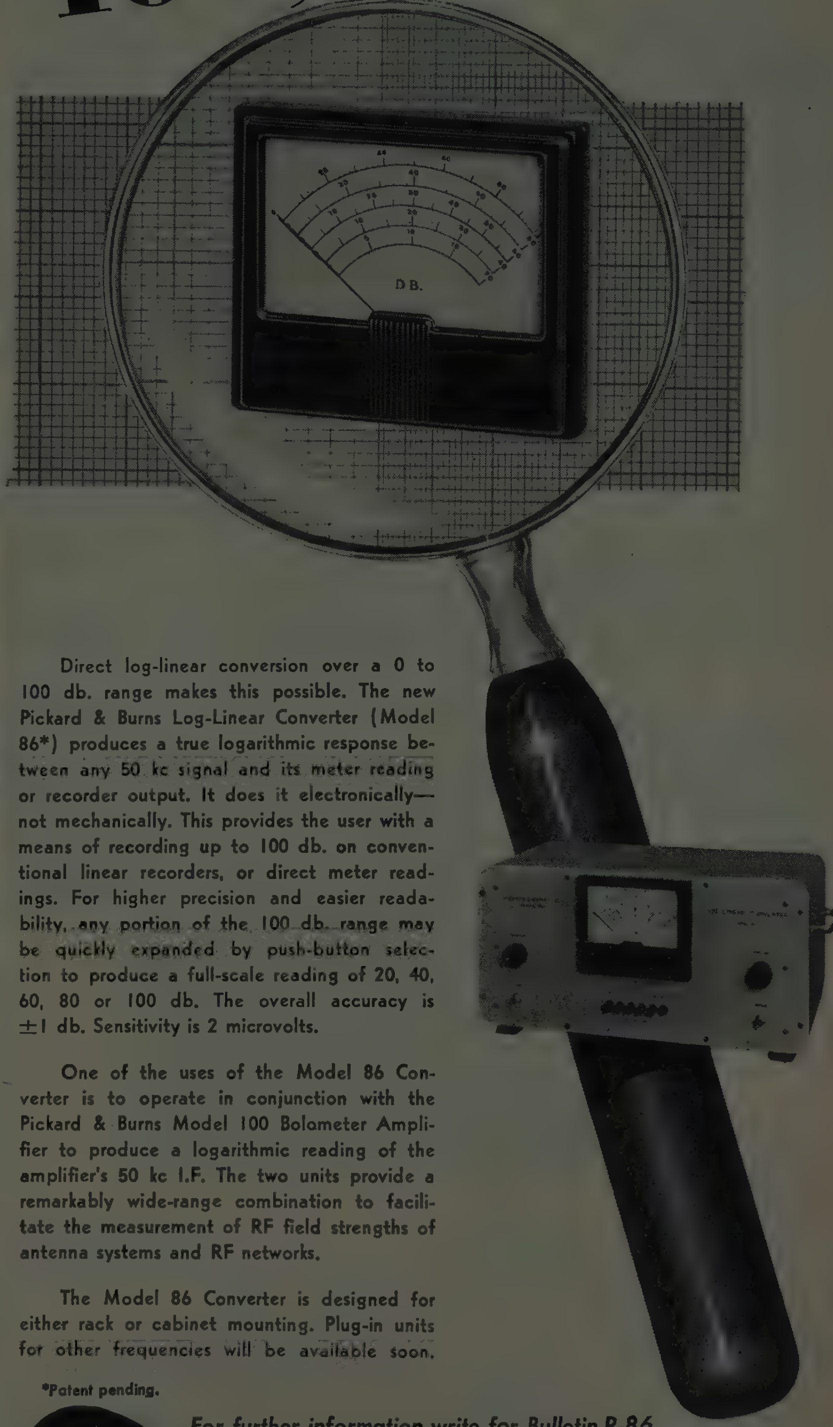
for further information write...

FASTER MELTING · POSITIVE FLUXING

ALPHA METALS, INC.
61 Water St., Jersey City 4, N. J.

Specialists **IN SOLDER** For Over 50 Years

NOW READ VOLTAGE RATIOS OF 100,000 to 1



Direct log-linear conversion over a 0 to 100 db. range makes this possible. The new Pickard & Burns Log-Linear Converter (Model 86*) produces a true logarithmic response between any 50 kc signal and its meter reading or recorder output. It does it electronically—not mechanically. This provides the user with a means of recording up to 100 db. on conventional linear recorders, or direct meter readings. For higher precision and easier readability, any portion of the 100 db. range may be quickly expanded by push-button selection to produce a full-scale reading of 20, 40, 60, 80 or 100 db. The overall accuracy is ±1 db. Sensitivity is 2 microvolts.

One of the uses of the Model 86 Converter is to operate in conjunction with the Pickard & Burns Model 100 Bolometer Amplifier to produce a logarithmic reading of the amplifier's 50 kc I.F. The two units provide a remarkably wide-range combination to facilitate the measurement of RF field strengths of antenna systems and RF networks.

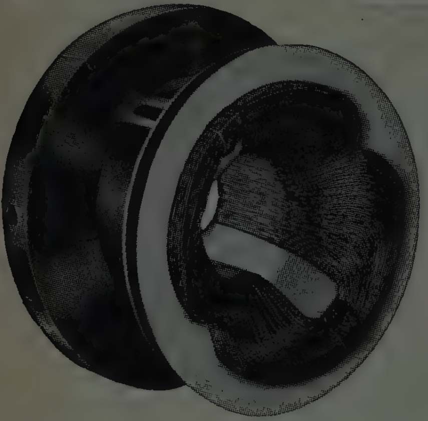
The Model 86 Converter is designed for either rack or cabinet mounting. Plug-in units for other frequencies will be available soon.

*Patent pending.

For further information write for Bulletin P-86

PICKARD & BURNS
INCORPORATED
24 Highland Avenue, Needham 94, Mass.

NOW DX 90° DEFLECTION YOKE
. . . an Accepted Standard of Major Manufacturers!



Yes, as leading TV manufacturers know . . . the DX 90° Deflection Yoke for 27″ receivers gives the ultimate in performance and compactness. Enclosed in an Underwriters' Approved Tenite case, this yoke assures a sharp, full-screen focus without use of pincushion magnets. Ingeniously designed for mass production on special equipment, it provides the attractive price and top quality major manufacturers demand. We invite your inquiry.

DEFLECTION YOKES . . . TOROID COILS . . . CRYSTALS
I. F. TRANSFORMERS . . . R. F. COILS . . . DISCRIMINATORS
SPEAKERS . . . TV TUNERS . . . ION TRAPS . . . TRANSFORMERS

DX COMPONENTS
"the heart of a good television receiver"
TRADE MARK

DX RADIO PRODUCTS CO.
GENERAL OFFICES: 2300 W. ARMITAGE AVE., CHICAGO 47, ILL.

NOISE and FIELD INTENSITY METER Model NF-105
(Commercial Equivalent of AN/URM-7)

ENGINEERING DATA

- Frequency range 20 MC to 1000 MC by means of three quick-change tuning heads. At least one RF amplifier with tuned input is employed for each tuning range.
- Easy frequency scanning by means of single tuning knob.
- Built-in impulse noise calibrator flat to 1000 MC (Output externally available).
- 4½″ logarithmic indicating meter reads carrier or true peak.
- Aural slideback operation.
- Input VSWR better than 1.2 to 1.
- Built-in regulated "A" and "B" supply.
- Complete line of accessories available.

Visit our booth at the Western IRE Show

edp EMPIRE DEVICES PRODUCTS CORPORATION
38-15 BELL BOULEVARD • BAYSIDE 61, NEW YORK

MANUFACTURERS OF
FIELD INTENSITY METERS • DISTORTION ANALYZERS • IMPULSE GENERATORS • COAXIAL ATTENUATORS • CRYSTAL MIXER

The Radio Engineering Show—1954 at KINGSBRIDGE Armory—New York City—March 22–25

Dedicated to the advancement of Engineering in Radio-Electronics, the annual Radio Engineering Show is an integral part of IRE's National Convention. It is a practical engineering exhibition of the newest manufacturing achievements in a highly technical industry, where knowledge of components, apparatus and instruments is the engineers' key to progress.

KINGSBRIDGE is the solution to three pressing problems confronting IRE after seven years in Grand Central Palace—now forced to move because the government has leased the Palace. KINGSBRIDGE serves these ways:

- It keeps a great show in a great city, central to largest audience of IRE member and non-member radio engineers.
- It lets the show grow, thus representing an ever-expanding industry.
- It helps "hold the line" on mounting exhibit costs and yet will provide a more beautiful and practical exhibition.

The following pages present KINGSBRIDGE as an exhibition hall, and describes the Radio Engineering Show for 1954, with facts for prospective exhibitors.

What Is IRE?

The Institute of Radio Engineers is the second largest engineering society in the United States. Organized in 1913, it now has 34,414 members with sections meeting in 76 cities. It is strictly a professional organization, with no company memberships or sponsors. It operates solely for the benefit of its members, as a non-profit corporation, advancing its industry by serving the individual engineer. This is done by the unrestricted exchange of engineering and scientific knowledge through the media of meetings and publications.

Meetings Make a Market-Place!

Manufacturers know that engineers "set the specs" for buying in a technical industry. The large IRE Meetings are a logical exhibit place for those products that the engineer possesses the knowledge to buy.

History of IRE Shows

The first Radio Engineering Show was held as a part of the IRE National Convention in 1930. These Shows continued successfully until stopped by wartime restrictions in 1942. But by 1945 the Government recognized the need for this exchange of information and the Show resumed with special permission. Since that time it has

(Continued on page 114A)

PRD

TYPE 904
VHF-UHF
NOISE
GENERATOR

FREQUENCY RANGE
(mc/sec): **10 to 1000**
NOISE FACTOR RANGE
(db): **20**
CHARACTERISTIC IMPEDANCE:
50 ohms (unbalanced)

This calibrated broadband noise source permits direct measurements of noise factors as high as 20 db for r-f amplifiers and receivers operating in the range from 10 to 1000 mc/s. Equipment is housed in an attractive metal cabinet.

Write for Catalog

Polytechnic
RESEARCH & DEVELOPMENT CO., Inc.
55 JOHNSON ST., BROOKLYN 1, N.Y.
West Coast:
741 N. Seward St., Hollywood 38, Cal.

Rapidly expanding programs in Microwave

SEMICONDUCTORS
COMPONENTS
MAGNETRONS
TR TUBES

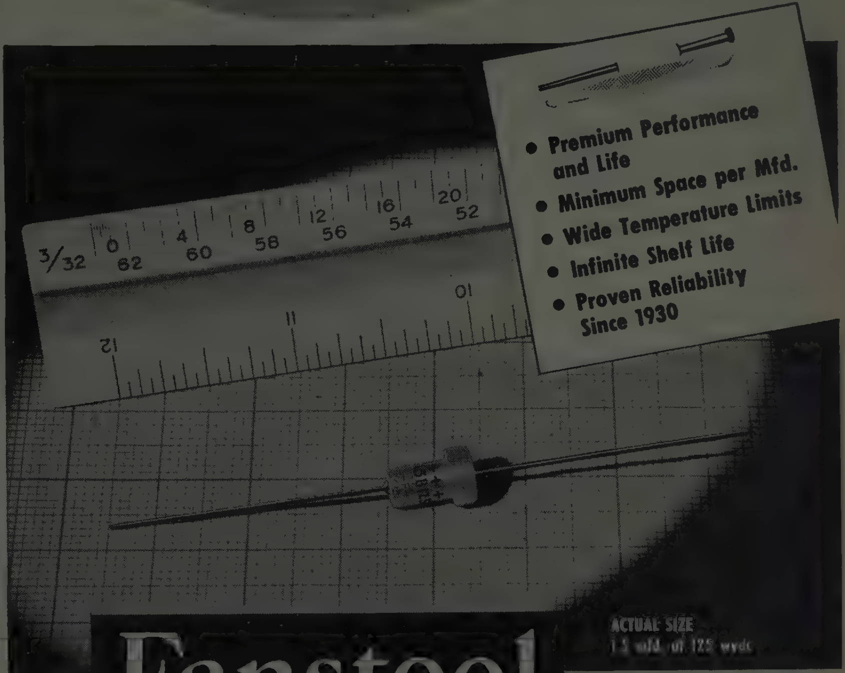
We need Senior and Junior Electronics Engineers and Physicists with background and interests in the above categories. Positions are available in Research, Design, Production Engineering. Previous experience in Microwave or UHF region desirable. Engineers now working on duplexers and mixers are especially suited to our needs.

Our young company offers tremendous growth opportunities along with paid vacations, free life insurance, free hospitalization for both employee and dependents. Write now for a descriptive brochure. Why not drop in during your vacation for an interview. Send resumé to "Director of Engineering" at address below:

μA Microwave Associates INC

A 50% affiliate of American Broadcasting-Paramount Theatres, Inc.
22 Cummington Street
Boston 15, Mass.

TANTALUM CAPACITORS **ARE BASIC**
in current electronic trend......

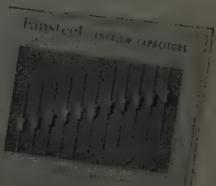

• Premium Performance and Life
• Minimum Space per Mfd.
• Wide Temperature Limits
• Infinite Shelf Life
• Proven Reliability Since 1930
ACTUAL SIZE
1.5 mfd. at 125 wvdc

Fansteel

TANTALUM CAPACITORS

Now, through the use of tantalum, new high standards of electrolytic capacitor performance are available. The tantalum oxide film is the most stable dielectric, chemically and electrically, yet discovered. As a result, Tantalum Capacitors offer advantages not found in any other electrolytic type—long life, space saving, wide temperature range, excellent frequency characteristics, no shelf aging.

Tantalum Capacitors are made by Fansteel and other leading capacitor manufacturers. Ask for current information bulletins on Fansteel Tantalum Capacitors.



FANSTEEL METALLURGICAL CORPORATION
NORTH CHICAGO, ILLINOIS, U. S. A.
Tantalum Capacitors... Dependable Since 1930

CAVITY FREQUENCY METER

NARDA Model 807

The ONLY precision meter available for the complete waveguide bandwidth of 12,400 to 18,000 megacycles.

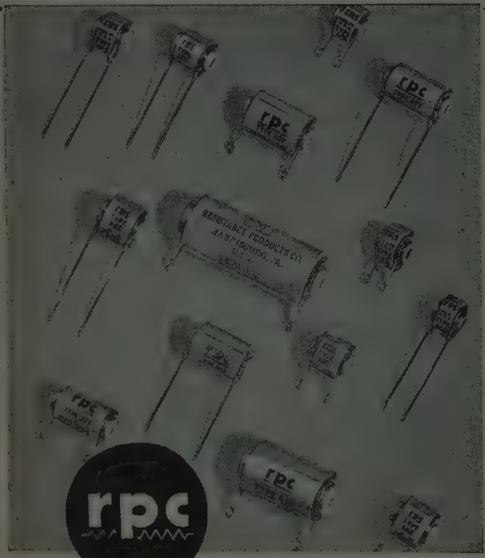
0.05% Absolute Accuracy — Precise Micrometer Drive — Calibration Data Conveniently Mounted — Loaded Q=5000— Electroformed Cavity

NARDA
Model 807

The NARDA Line
INCLUDES:
Hi-Power Impedance Meters • 2400-10,200 MC Frequency Meter • Pulsed R-F Signal Generators • Microwave Mixers and Detectors • Field and Laboratory Test Eqpmt. • Electronic and Microwave Consulting, Design and Testing.
Custom built Electronic and Microwave Equipment.

Send for additional data and prices on this and other Narda Models.

NARDA NASSAU RESEARCH & DEVELOPMENT ASSOCIATES INCORPORATED
66 MAIN STREET MINEOLA, NEW YORK
Subsidiary of PEERLESS INSTRUMENT CO., INC. GARDEN CITY 3-3570



Leading Manufacturers Rely on RPC for Quality and Quantity in Precision Resistors!

Within a few years RPC has attained a position of leadership in the manufacture of precision wire wound resistors. This is no accident. It is the result of STRINGENT *control of quality*—use of the *finest available materials*—test equipment and standards that are matched only by a few outstanding laboratories.

RPC Precision Resistors meet Government specifications. They are acceptable for all types of equipment—test instruments, electronic computers and scientific equipment. Requirements of Jan-R-93, Mil-R-93A are fully met. Advanced methods of production have made possible large or small orders at reasonable cost with prompt delivery. Write for catalog and helpful information about RPC's resistors.

RESISTANCE PRODUCTS CO.
714 Race St. Harrisburg, Penna.
Precision Wire Wound—High Voltage—High Megohm—High Frequency—Hermetically Sealed—QUALITY RESISTORS

RPC Type	Government Specification		Dimensions		Resistance Ohms			Watts	
	Jan-R-93	Mil-R-93A	Length	Diameter	Min.	Max. with Low T.C. Alloy .0015 Dia.	Max. with Low T.C. Alloy .001 Dia.	Jan or Mil	Comm'l
AFB*	RB10	RB15	15/32	17/32	0.1	.160 meg	.650 meg	.25	.5
AGB*	RB10	RB15	15/32	5/8	0.1	.235	1.0	.25	.5
AFC*	RB11	RB16	5/8	17/32	0.1	.225	1.0	.33	.5
AGC*	RB11	RB16	5/8	5/8	0.1	.330	1.5	.33	.5
AFF*	RB12	RB17	1	17/32	0.1	.475	2.0	.5	1
AGF*	RB12	RB17	1	5/8	0.1	.700	3.0	.5	1
AJS	RB13	RB18	1-9/32	11/16	0.1	1.25	5.0	.5	1
ALP	RB14	RB19	2-1/16	13/16	0.1	2.5	10.0	1	2

*Note—Can be furnished with 1-1/2" long 20 gauge tinned wire leads instead of lug terminals. Suffix "W" after type denotes wire leads.

(Continued from page 112A)

sky-rocketed as shown by the attendance tables on page 120A. Today it ranks in size with the larger trade shows, and is widely considered as a "radio-electronic industry show."

What and How to Exhibit!

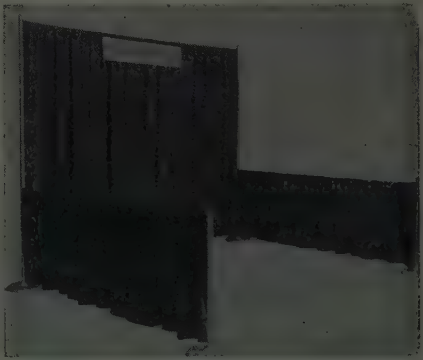
However, it can be a serious mistake to plan activities and an exhibit as if this is an ordinary industry show! The 30,000 engineers who will come up to KINGSBRIDGE expect to see informative, technical exhibits, and to talk specific problems with engineers.

Your exhibit should have these objectives, to pay-off in profitable sales:

1. To exhibit your newest engineering products, services or improvements to products previously shown.
2. To establish direct contacts between visiting engineers and your own engineers for idea exchange.
3. To produce inquiries and sales.
4. To demonstrate hard-to-sell items, or those requiring good and controlled exhibit conditions.
5. To show how wide your lines, and flexible or big your organization.
6. To "sound out" the market for new or needed products you can make. To discover new markets, or new uses for established products.
7. To crystallize into buying, the interest in products your advertising has created, by "product presentation and demonstration."

Very respectfully, we offer these "don'ts" as likely to result negatively:

1. Exhibits for employment purposes are contrary to Show Rules.
2. The billboard, or strictly institutional exhibit creates disappointment because visitors want to see products.
3. Do not underman your exhibit. Be sure to staff it with engineers.
4. Do not plan to pick up your prospects on the Show Floor to take to the hotel to talk business. KINGSBRIDGE is remote for this purpose, and we recommend against an exhibit if this is your method. In this Show the exhibit itself is the place for business. Engineers prefer that method, as KINGSBRIDGE is certain to prove.



New Booth Saves Money

A new booth has been designed for us by the Ivel Construction Corporation, our official decorators. The backwall and

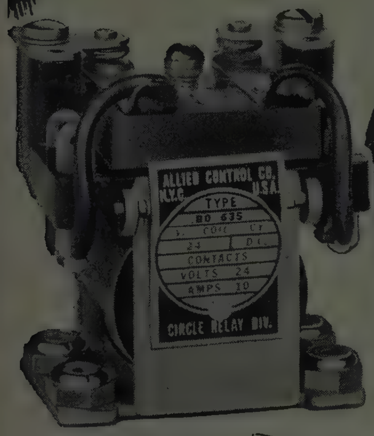
(Continued on page 118A)

24 HOUR
DELIVERY
FROM STOCK!

RELAYS

Our stock of more than a million relays — in over a thousand different types — is the world's largest. Don't delay your production for want of large or small quantities of relays of any type.

Telephone, wire or write for quotations.


ALLIED CONTROL CO.
N.Y.C. U.S.A.
TYPE
BO 635
CONTACTS
VOLTS 24
AMPS 10
CIRCLE RELAY DIV.

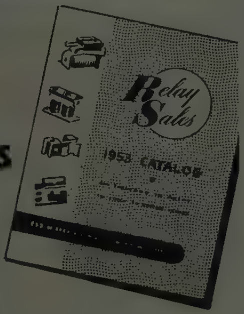
NEW AND MORE COMPREHENSIVE

1953

RELAY SALES CATALOG

NOW READY

Be sure to send for your copy

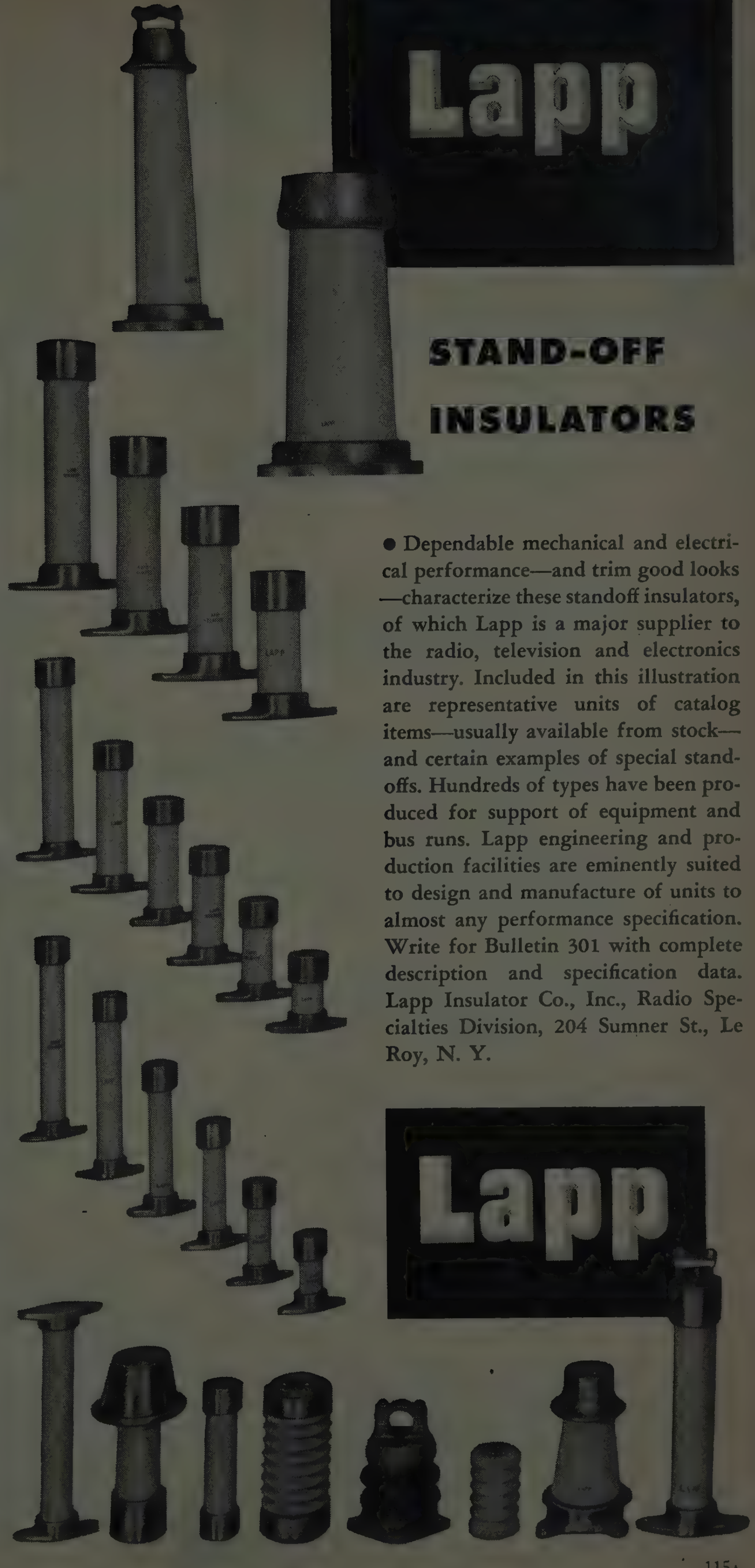

Relay Sales
1953 CATALOG

Relay Sales

Telephone
SEeley 8-4146

833 W. CHICAGO AVE.
DEPT. 7, CHICAGO 22, ILL.

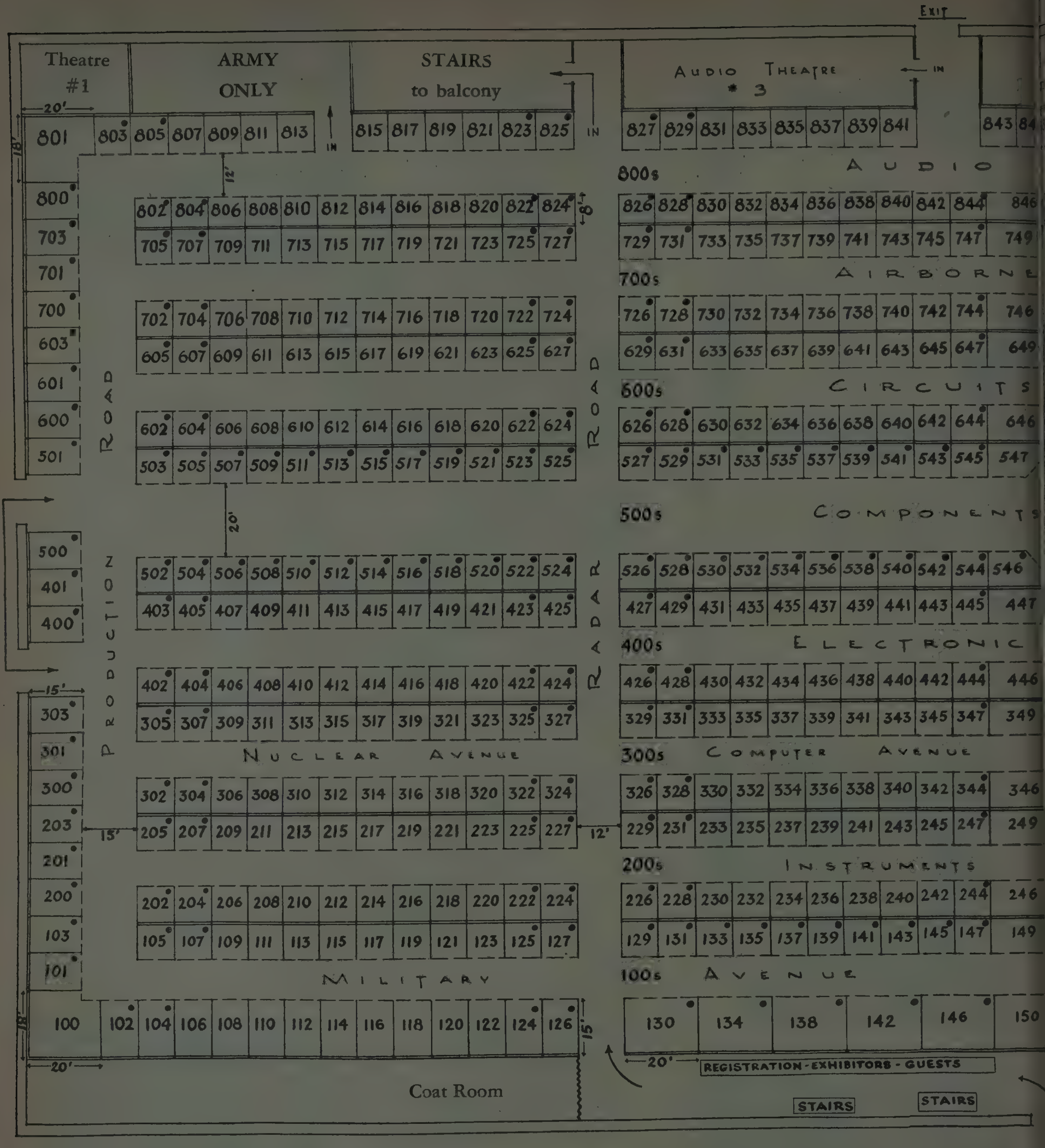
Lapp



STAND-OFF INSULATORS

● Dependable mechanical and electrical performance—and trim good looks—characterize these standoff insulators, of which Lapp is a major supplier to the radio, television and electronics industry. Included in this illustration are representative units of catalog items—usually available from stock—and certain examples of special standoffs. Hundreds of types have been produced for support of equipment and bus runs. Lapp engineering and production facilities are eminently suited to design and manufacture of units to almost any performance specification. Write for Bulletin 301 with complete description and specification data. Lapp Insulator Co., Inc., Radio Specialties Division, 204 Sumner St., Le Roy, N. Y.

Lapp


EXIT
Theatre #1
ARMY ONLY
STAIRS to balcony
AUDIO THEATRE # 3
IN
20′
18′
12′
8′
20′
15′
15′
12′
18′
20′
15′
20′
801 803 805 807 809 811 813
815 817 819 821 823 825
827 829 831 833 835 837 839 841
843 84
800s AUDIO
802 804 806 808 810 812 814 816 818 820 822 824
826 828 830 832 834 836 838 840 842 844 846
705 707 709 711 713 715 717 719 721 723 725 727
729 731 733 735 737 739 741 743 745 747 749
700s AIRBORNE
702 704 706 708 710 712 714 716 718 720 722 724
726 728 730 732 734 736 738 740 742 744 746
605 607 609 611 613 615 617 619 621 623 625 627
629 631 633 635 637 639 641 643 645 647 649
600s CIRCUITS
602 604 606 608 610 612 614 616 618 620 622 624
626 628 630 632 634 636 638 640 642 644 646
503 505 507 509 511 513 515 517 519 521 523 525
527 529 531 533 535 537 539 541 543 545 547
500s COMPONENTS
502 504 506 508 510 512 514 516 518 520 522 524
526 528 530 532 534 536 538 540 542 544 546
403 405 407 409 411 413 415 417 419 421 423 425
427 429 431 433 435 437 439 441 443 445 447
400s ELECTRONIC
402 404 406 408 410 412 414 416 418 420 422 424
426 428 430 432 434 436 438 440 442 444 446
305 307 309 311 313 315 317 319 321 323 325 327
329 331 333 335 337 339 341 343 345 347 349
NUCLEAR AVENUE
300s COMPUTER AVENUE
302 304 306 308 310 312 314 316 318 320 322 324
326 328 330 332 334 336 338 340 342 344 346
205 207 209 211 213 215 217 219 221 223 225 227
229 231 233 235 237 239 241 243 245 247 249
200s INSTRUMENTS
202 204 206 208 210 212 214 216 218 220 222 224
226 228 230 232 234 236 238 240 242 244 246
105 107 109 111 113 115 117 119 121 123 125 127
129 131 133 135 137 139 141 143 145 147 149
MILITARY
100s AVENUE
100 102 104 106 108 110 112 114 116 118 120 122 124 126
130 134 138 142 146 150
800 703 701 700 603 601 600 501
500 401 400
303 301 300 203 201 200 103 101
PRODUCTION ROAD
RADAR ROAD
FREIGHT ENTRANCES
Coat Room
REGISTRATION - EXHIBITORS - GUESTS
STAIRS
STAIRS

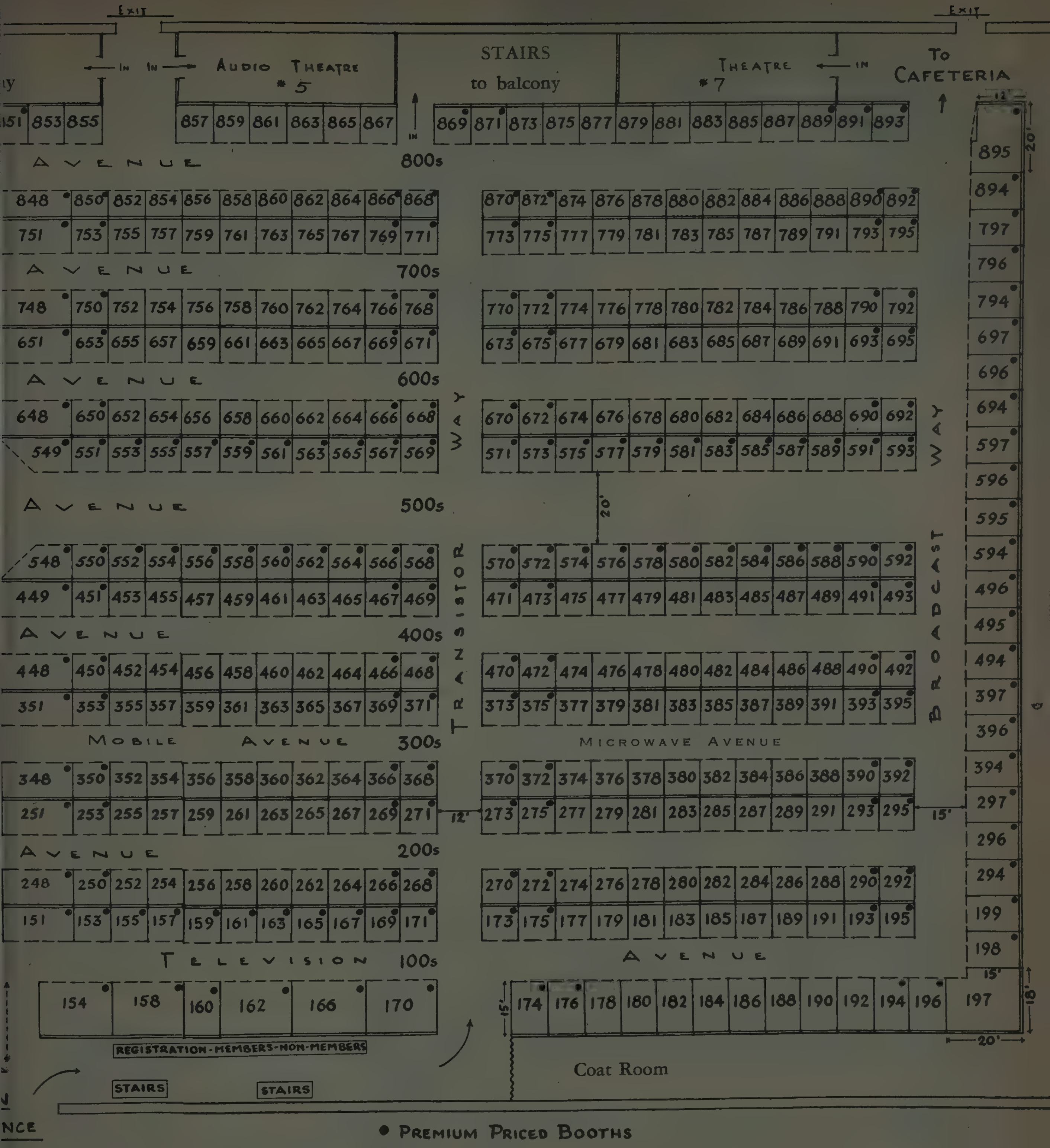
Note:
Water & Gas May Be Supplied
Along North & South Walls

Kingsbridge A

Booth Sizes: All units have a 10′ backwall and frontage, double booths 20′. Most units are 10′×10′ or 100 sq. ft. Exceptions: Units 802 to 892 are only 8′ deep, 80 sq. ft. The East, South and West wall units are 15′ deep except 130–170 which are 12′ deep. Theater units differ from this plan, please ask us.

Positions and Rates: All booths shown with black dot in upper right corner are preferred positions, those without are regular. Rates:

	Preferred	Regular
Price per sq. ft.	$4.40	$3.30
6 Time Proceedings advertisers	$4.00	$3.00
12 Time Proceedings advertisers	$3.60	$2.70


EXIT
EXIT
IN
IN
AUDIO THEATRE #5
STAIRS to balcony
THEATRE #7
IN
To CAFETERIA
IN
12'
20'
ly
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869 871 873 875 877 879 881 883 885 887 889 891 893
895
AVENUE 800s
848 850 852 854 856 858 860 862 864 866 868
870 872 874 876 878 880 882 884 886 888 890 892
894
751 753 755 757 759 761 763 765 767 769 771
773 775 777 779 781 783 785 787 789 791 793 795
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AVENUE 700s
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770 772 774 776 778 780 782 784 786 788 790 792
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673 675 677 679 681 683 685 687 689 691 693 695
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AVENUE 600s
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AVENUE 500s
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AVENUE 400s
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448 450 452 454 456 458 460 462 464 466 468
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351 353 355 357 359 361 363 365 367 369 371
373 375 377 379 381 383 385 387 389 391 393 395
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MOBILE AVENUE 300s
MICROWAVE AVENUE
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370 372 374 376 378 380 382 384 386 388 390 392
394
251 253 255 257 259 261 263 265 267 269 271
12'
273 275 277 279 281 283 285 287 289 291 293 295
15'
297
AVENUE 200s
296
248 250 252 254 256 258 260 262 264 266 268
270 272 274 276 278 280 282 284 286 288 290 292
294
151 153 155 157 159 161 163 165 167 169 171
173 175 177 179 181 183 185 187 189 191 193 195
199
TELEVISION 100s
AVENUE
198
TRANSISTOR WAY
BROADCAST WAY
154 158 160 162 166 170
15'
174 176 178 180 182 184 186 188 190 192 194 196 197
15'
18'
20'
REGISTRATION-MEMBERS-NON-MEMBERS
Coat Room
STAIRS
STAIRS
NCE
● PREMIUM PRICED BOOTHS

nory NEW YORK CITY

Backwall and Height: Rental includes 8′ high blue drapes and 30″ high siderail drapes supported on polished aluminum stanchion posts. All 1954 exhibits are limited to 8′ high. A name card 9″×44″ and blue broadfelt rug is also supplied. Balcony is 14′7″ and arched ceiling rises to 120′ at center of hall.

System of Assignment: New exhibits are accepted only on a waiting list basis. Past exhibitors are assigned on date of order after offer. Offers are made floor by floor to exhibitors in Palace in accordance with the Exhibitor's Committee's recommended plan. A list of assigned booths sent on request.

D.C. MILLIAMPERES
BURLINGTON

Burlington

RUGGEDIZED
HERMETICALLY SEALED
INSTRUMENTS

LABORATORY APPROVED
Under MILITARY SPECIFICATION MIL-M-10304

- Will withstand Hi-Impact Shock Tests, extensive vibration and tumbling requirements, extreme thermal shock, humidity and temperature tests.
- Glass to metal type seal with molded vulcanized rubber terminal connections capable of carrying 30 amps for self-contained units.
- Magnetically shielded.
- Black satin onodized aluminum bezel.
- D'Arsonval permanent magnet type movement—for all D.C. Ranges.
- Available in 2½" and 3½" round case types.
- Guaranteed one year against defective workmanship and materials.

BURLINGTON INSTRUMENTS ALSO AVAILABLE IN HERMETICALLY SEALED OR BAKELITE CASES IN 1½" THROUGH 4½" SIZES

BURLINGTON INSTRUMENT COMPANY
DEPT. I-83, BURLINGTON, IOWA

Specification Coils

—for every requirement—radio, FM, TV and Government Applications!

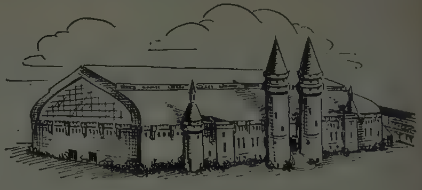
Including Universal, Bank Wound, Universal Progressive and Solenoid. All are precision-built to highest engineering standards and conform exactly to specifications. For uniform high quality, prompt delivery and economical unit costs, specify coils by Fugle-Miller. Radio, TV and JAN specifications are a specialty. Phone, wire or write for quotations.

ADDRESS INQUIRIES TO DEPT. P-8

FUGLE-MILLER LABORATORIES
MAIN STREET, METUCHEN, NEW JERSEY
Telephone: Metuchen 6-2245

(Continued from page 114A)

matching siderail drapes are of glossy middle-blue, supported by polished aluminum rail stanchions and capped with black plastic, streamlined knobs. A free handlettered sign, 44″×9″ and blue broadfelt rug completes the IRE package.



End of the Backwall Race

As voted in the Exhibitors Advisory Meeting the KINGSBRIDGE Show adopts an 8′ height limit throughout the entire show and all areas. This brings to an end the race between outside wall exhibitors to higher and higher wall drapes which were costly without any real effect in selling, or informing engineers. (See rules on page 122A.)

Faster Trucking In and Out

Trucks can be unloaded and loaded right on the exhibition floor, in front of the exhibit. There is no limit to the floor load. Up to 20 trucks can be handled on the floor at one time, and fork-lift unloading will be used. This promises to end the loading platform bottleneck which has made moving in and out of Grand Central Palace such a problem.

If exhibitors will cooperate with the Exhibits Manager on the schedule which will be issued in the Exhibitor's Guidance Book, costs can be very greatly reduced, and delays ended. All box storage is in the building, under cover and unlimited. The service is supplied free, up to the riggers' weight limits.

Transportation

A bus free to exhibitors and registered guests will be operated leaving the Waldorf-Astoria and KINGSBRIDGE every ten minutes. The Allied Line Buses are luxury coaches. Time: 40 minutes.

Subways

The map on page 126A illustrates the subways serving KINGSBRIDGE. From Grand Central Station it is 30 minutes to a stop on the corner of the building. The "D" trains stop nearby.

Parking

Special show-time parking is being arranged on 195th and adjacent streets. Undercover parking in four nearby garages can be arranged for 200 exhibitor's cars.

The Cafeteria

An attractive cafeteria seating 505 will be operated in the Officers' Dining Room, directly below the exhibition floor. Five choices in good, hot meat dinners will be offered.

(Continued on page 120A)

WHERE RESISTANCE TO HIGH AND LOW TEMPERATURES IS VITAL

Surflene

INSULATED HOOK-UP WIRE

Resists

- HEAT
- FUNGI
- ABRASION
- CHEMICALS

"Surflene", extruded monochlorotrifluoroethylene, has high insulation resistance, dielectric strength and outstanding resistance to heat, abrasion, most chemicals and concentrated acids, including fuming nitric acid. It is non-inflammable and inert to fungi. It is especially designed for hermetically sealed and water-proof equipment and for high temperatures encountered in power supply and continuous duty apparatus. Also available in multi-conductor cables.

"Surflene" is available in thirteen colors — red, orange, yellow, pink, light and dark green, blue, gray, tan, brown, black, white and clear.

Write our Engineering Service TODAY for technical assistance.

Surprenant MFG. CO.
199 Washington St. Boston 8, Mass. Plant—Clinton, Mass.
Engineered Wire and Cable for the Electronic and Aircraft Industries

CANNON PLUGS *for hermetic sealed applications*

KH

RKH

HERMETIC SEALED Type RKH Plugs and KH Receptacles mate with their corresponding Cannon RK and K standard fittings. The basic construction of fused vitreous insulation around the contacts is same as GS type. Shell materials and finish are likewise similar. Various types of flange or hex-bulkhead styles may be made to order.

Refer to KH-1 Section in K Bulletin.

SUB-MINIATURE receptacles of the new Cannon "U" Series are used on miniature switches, relays, transformers, amplifiers, and other sealed components, requiring a true hermetic seal or a connector of sub-miniature size with performance superiority.

"U" plugs have a steel shell and "SILCAN*" insulator, cable relief and moisture resistant sleeve.

Bayonet-type locking means prevents vibration failure. Rated 1700v. d.c.; 5a. Available in 3, 6, and 12 contact arrangements with one plug style and two receptacles.

**Cannon Electric's special silicone resilient material.*

Refer to U-2 Bulletin

GS02

GS06

GS Types mate with standard AN(MIL) types. These highly successful hermetically sealed plugs (GS06) and receptacles (GS02) pioneered this field and are top quality fittings. Fused vitreous insulation provides a true hermetic seal for relays, position indicators, etc. Shells are steel, finished in cadmium plate and bleached Iridite; coupling nut on plug is natural finish Dural. Eyelet or solder pot terminals.

Built to resist thermal shock, —300°F. to +600°F., surpassing MIL Spec. GS02 Types will withstand operation temperatures 400°F. to 600°F., and pressures as high as 200 to 900 psi; specials to 7500 psi. GS Types approximate AN voltage and current ratings. Wide range of AN layouts available.

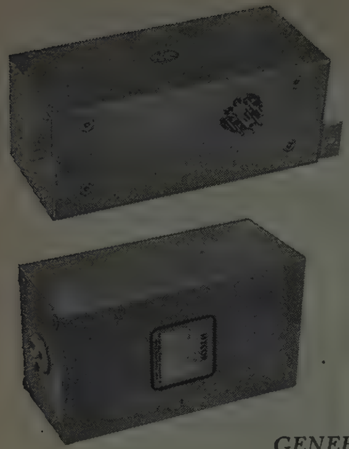
See GS-3 section in AN-8 Bulletin for details.

COMING: TYPE "DH" HERMETIC SEALED CONNECTORS SIMILAR TO PRESENT DA-15P

CANNON ELECTRIC
Since 1915

Factories in Los Angeles, Toronto, New Haven, Benton Harbor. Representatives in principal cities. Address inquiries to Cannon Electric Co., Dept. H-377, P.O. Box 75, Lincoln Heights Station, Los Angeles 31, Calif.

HYCOR



Telemetering Filters

FEATURES . . .

- HYCOR telemetering filters have excellent characteristics due to the use of high "Q" toroid inductor elements. The filters may be used in low level circuits with negligible hum pickup resulting.
- In addition, only the finest capacitors are employed to assure stability.
- Available in standard RDB frequencies.

GENERAL SPECIFICATIONS
IMPEDANCE 500/500 OHMS

TYPE	BANDWIDTH	ATTENTUATION	FREQUENCY RANGE	PRICE
1500	±7½%	−3 db or less	400 cps to 14.5 kc.	$ 55.00
	±20%	−30 db or more		
4300	±7½%	−3 db or less	400 cps to 960 cps	82.50
	±20%	−40 db or more	1300 cps to 14.5 kc.	75.00
4000	±7½%	−3 db or less	400 cps to 960 cps	105.00
	±15%	−45 db or more	1300 cps ot 14.5 kc.	95.00
	±15%	−3 db or less	22 kc. to 70 kc.	
	±28%	−45 db or more		

Other frequencies and impedances available on request

11423 VANOWEN ST., NORTH HOLLYWOOD, CALIFORNIA **HYCOR** Company, Inc. SUnset 3-3860

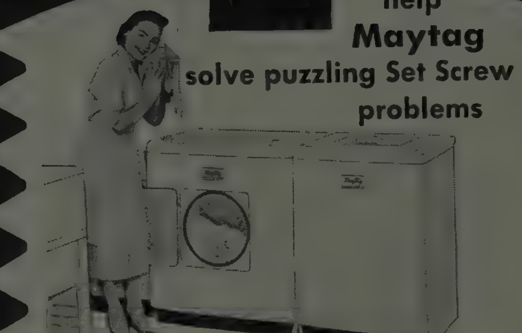
Manufacturers of Precision Resistors, Toroid Inductors and Electric Wave Filters

REPRESENTATIVES:
Jack Beebe, 5707 W. Lake Street, Chicago, Illinois
George E. Harris & Co., Box 3005, Municipal Airport, Wichita, Kansas
Marvin E. Nulsen, 5376 E. Washington St., Indianapolis 19, Indiana
Burlingame Associates, 103 Lafayette Street, New York City
Canada: John R. Tilton, 1166-A Lake Shore Road, Long Branch, Ontario

For further information contact your nearest Hycor representative or write for Bulletin TF

SETKO SET SCREWS

help Maytag solve puzzling Set Screw problems



Maytag helps housewives solve puzzling laundry problems, with its famous Maytag Washers and this new Automatic Clothes Dryer. When Maytag has puzzling set screw problems, it comes to Set Screw & Manufacturing Co., as nearly 1500 other manufacturers have done ever since 1935. Today Setko Set Screws are used in all Maytag washers. If you have a puzzling set screw problem, regarding size, delivery, unusual or difficult application, we can probably help you solve it with Setko, Zip-Grip* Self-Locking or the new Nu-Cup Set Screws. Write us, describing your set screw application, and we will send you Free Test Samples and Zip-Grip* Demonstrator.

*Pat. Pending

228

Set Screw & Mfg. Co.

FOR FREE CATALOG

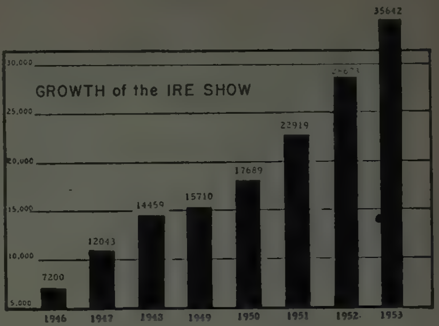
Simply write your name and address on margin of this page, tear out and mail today.

422 Main St., Bartlett, Ill. (Chicago suburb)

We Specialize in Solving Puzzling Set Screw Problems

(Continued from page 118A)

Attendance at the Radio Engineering Show



A detailed analysis of attendance at the past shows, comprised of five tables is available on request. It classifies visitors by type of business and job titles.

Golden Rule

When nearly 500 different exhibitors cooperate to present a huge exhibition to more than 30,000 visiting engineers, rules are a necessary part of "order." The spirit behind these rules is "the Golden Rule."

Purpose and Rules

Purpose: The purpose of the Radio Engineering Show is to present material pertinent to the technical interests of radio and electronic engineers. In order to assure that the exhibition will retain a technical and engineering character, the right to limit attendance to qualified engineers, revise floor plans, and shift exhibits, and to disapprove any exhibit is reserved to The Institute of Radio Engineers and the Exhibits Manager. Exhibits having only lay interest are not permitted. Employee solicitation is restricted.

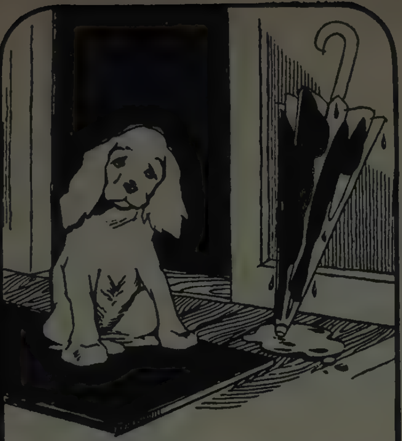
Rental is $4.40 per square foot for preferred space and $3.30 for regular space. Theatres and some non-standard booths have "unit" prices. Rate of $4.00 and $3.00 is allowed firms using six product advertisements worth $500 or more in the PROCEEDINGS of the IRE, and $3.60 and $2.70 for 12 time "product" advertisers during 1953 or in the 12 months preceding the Show.

Payment: A 50% deposit is due with reservations and the balance by January 2, 1954. Cancellations by exhibitors will be accepted only at the discretion of the Institute and deduction for expenses due to cancellation will be made against refunds. No space assignment is binding to the IRE unless the 50% deposit has been made, and IRE contract signed.

Make Checks Payable To: The Institute of Radio Engineers

Space assigned to an exhibitor may be transferred by Exhibits Manager to balance the Show against congestion; to avoid confusion in firm names; to solve competitive conditions, or similar reasons, and if payment schedule is not maintained. An exhibitor may not share or sublet space to another party. One exhibitor may not exhibit the named end product of another exhibitor without permission of that exhibitor in writing.

(Continued on page 122A)



MOISTURE PROBLEMS ARE ENDED . . .

WHEN YOU SPECIFY

Standard Piezo CRYSTALS

Continuing research, "up-to-the-minute" manufacturing methods and constant laboratory and field testing assure you of truly accurate and dependable performance under all types of unfavorable conditions, when you insist on Standard Piezo Crystals for your equipment. Send today for our new, free illustrated catalog.

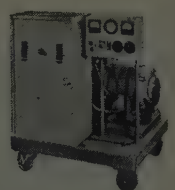
Standard Piezo COMPANY

CARLISLE, PENNA.

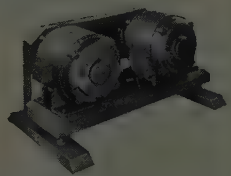
"packaged" *high frequency* power supplies *for ground installations*

American Electric supplies these complete, *"packaged unit"* power supplies for *all* high frequency requirements. Noted for their rugged, reliable performance, the amazingly low maintenance factor of American Electric alternators results from a unique and exclusive design principle: *a rotating inductor without coils; without springs, slip rings or brushes!* Nothing to wear out, nothing to service . . . *as maintenance-free as its two sealed ball bearings!*

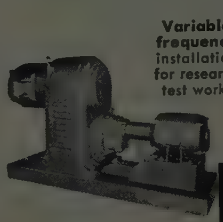
Semi-portable, skid-mounted missile launching power supply completely weather protected.



Portable, caster-mounted 400 cycle installation for production test equipment.



Stationary installation, vee belt drive, resilient mounted for laboratory h.f. test equipment.



Variable frequency installation for research test work.

Manufacturers Also of Miniature A.C. (All Frequencies) Electric Drive Motors, Blowers and Fans

Correct Power Supply for Every Installation

Portable, semi-portable or stationary types, open models or completely enclosed for weather protection. Caster or pneumatic tire mounts, skid mounts and resilient rubber mounts on stationary types.

Wide Frequency Ranges

Fixed Frequencies from 250 cycles to 2400 cycles (up to 4000 cycles in the lower ratings).

Variable Frequencies from 380 cycles to 1200 cycles and 1200 cycles to 2400 cycles.

Excellent Voltage Regulation: Standard ± 1% to as low as ± .5% depending upon choice of drive. Electronic regulators or magnetic amplifier regulators supplied.

Motor Drives—Common shaft, direct connected, Vee belt or positive, no-slip timing belt types. Variable speed on variable frequency models.

Low Harmonic Content
Less than 2% on single phase.
Less than 1% on three phase.
Exceedingly low harmonic content results directly from alternator design without use of filters.

Output Ranges
single phase—500 watts to 15 KVA
three phase—500 watts to 30 KVA
(outputs up to 75 KVA available in other alternator designs.)

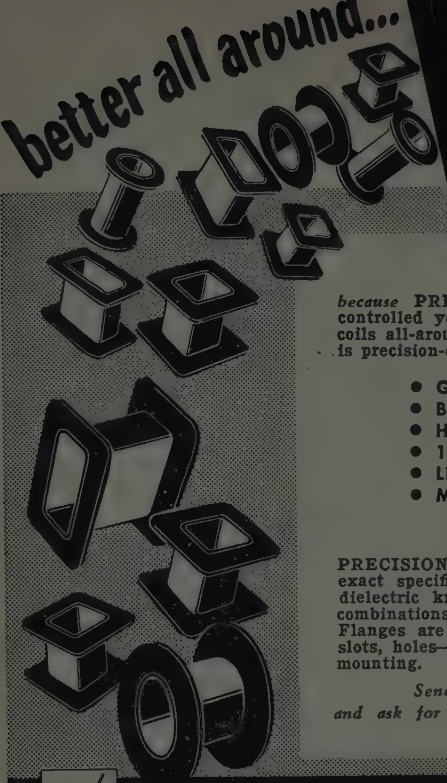
WHATEVER YOUR GROUND POWER SUPPLY REQUIREMENTS
Ask American Electric for quotation

NOW! Ask for Details!
American Electric Auxiliary Airborne
MISSILE POWER SUPPLIES

AMERICAN ELECTRIC MOTORS
TRADE MARK

4811 Telegraph Road,
Los Angeles 22,
California

Engineering Representatives: TRAVCO ENGINEERING CO.
Silver Spring (Md.), Chicago, New York, Los Angeles



better all around...

Precision
COIL BOBBINS

because PRECISION Coil Bobbins are quality controlled you can be sure they'll improve your coils all-around. Every PRECISION Coil Bobbin is precision-engineered to give you:

- Greater Insulation
- Better Heat Dissipation
- Higher Moisture Resistance
- 15 to 20% Greater Strength
- Lighter Weight
- More Winding Space

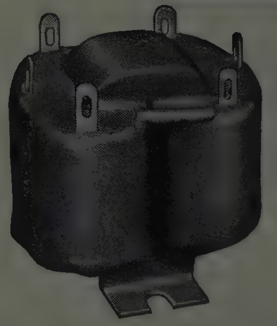
All at low unit cost!

PRECISION Coil Bobbins can be designed to your exact specifications . . . any size or shape in dielectric kraft or fish paper, cellulose acetate, combinations or phenol impregnated materials. Flanges are supplied plain or fitted with leads, slots, holes—flat, recessed or embossed to fit any mounting.

Send specifications for free sample and ask for new Arbor list of over 2000 sizes.

PRECISION

PRECISION PAPER TUBE CO.
2051 W. Charleston St. Chicago 47, Ill.
Plant No. 2, 79 Chapel St., Hartford, Conn.
Also Mfrs. of PRECISION Paper Tubes



MAKING TRANSFORMERS IS OUR BUSINESS

• For more than 35 years Acme Electric transformers have become components of all types of electrical and electronic equipment. The vast technical experience accumulated during this time is now available to west coast manufacturers through our Los Angeles branch.

ACME ELECTRIC CORPORATION
MAIN PLANT: 448 Water Street • Cuba, N. Y.
West Coast Engineering Laboratories: 1375 W. Jefferson Blvd. • Los Angeles, Calif.
In Canada: ACME ELECTRIC CORP. LTD. • 50 North Line Rd. • Toronto, Ont.

Acme ACME Electric
TRANSFORMERS

(Continued from page 120A)

Registration: The Radio Engineering Show is an integral part of the IRE National Convention. Visitors to the show, and exhibits personnel must register according to the rules of the convention. Exhibitors bringing in guests not in compliance with registration requirements will be billed for the registration of such guests. It is a "closed" show. No admission to minors under 18 years old unless at all times accompanied directly by an adult. Exhibitor's children may not "roam" the Show.

Sound: The use of head sets or oscilloscopes is recommended. For obvious reasons loud-speaking sound displays are not permitted. This rule will be rigidly enforced. Record changers may be shown in mechanical operation, but played only in sound theaters. Such special rooms for sound demonstrations are for rent. Exhibitors are not permitted to use public address systems in open exhibits; or demonstrations which when in action raise the sound level by 8 db, or are harmful to their neighbors.

Visuals: Small-sized motion picture or slide-film screens for showings are desirable, if positioned so as not to crowd aisles. Flashing or glaring light exhibits are not permitted, because they are harmful to the eyes of attendants or neighboring booths. Some animated displays are allowed.

Soliciting—Samples—Prizes: No soliciting for business shall be permitted in aisles or in other exhibitors' booths. Samples, catalogues, pamphlets, publications, etc., may be distributed by exhibitors *only within their own booths*. No exhibit will be permitted to conduct any prize drawings, awards for signing of names and addresses, or other extreme promotional stunts, etc., without first obtaining written permission from the IRE. Distribution of envelopes or bags for the accumulation of manufacturers' literature is prohibited.

Fire Laws: Federal, State, and City Fire Laws must be strictly observed. Cloth decorations must be flame-proof. Wiring must comply with fire department and underwriters rules. Smoking in exhibits may be forbidden. Crowding will be restricted. Aisles and fire exits cannot be blocked by exhibits. No decorations of paper, pine boughs, leafy decorations or tree branches are allowed. In New York City a certificate of flame-proofing of drapes is required (not more than one year old). Acetate and most rayon drapes are not fireproof.

Hours: In their own best interests, and for security, exhibitors are required to man their booths during all open hours. They may not dismantle before the show closes. Hours may be adjusted by the Exhibits Manager and Committee. Hours generally have been: Monday 11 A.M. to 9 P.M.; Tuesday 9:30 A.M. to 9 P.M.; Wednesday 9:30 A.M. fo 6 P.M. (Banquet Night) and Thursday 9:30 A.M. to 9 P.M.

Height Limit 8 Feet: All exhibits, backwalls and decorations are limited to 8 feet high; except actual radio-electronic equipment, for which permission in writing must be obtained from the Show Manager if it exceeds this height.

(Continued on page 124A)

C O N S T A N T

RESEARCH

and

DEVELOPMENT

HAS ENABLED US TO PRODUCE ONE OF THE FINEST COMPUTERS AVAILABLE TODAY.

AMPLIFIER GROUP, TYPE 16-31C—provides 28 contact stabilized operational amplifiers for use as summers, differentiators, integrators, and inverters. Also in the cabinet are all necessary power supplies and a complete test panel.

MULTIPLIER GROUP, TYPE 16-31L—is a servo-mechanical multiplier and incremental function generator. There are 20 channels, each of which is capable of multiplying four variables by a fifth.

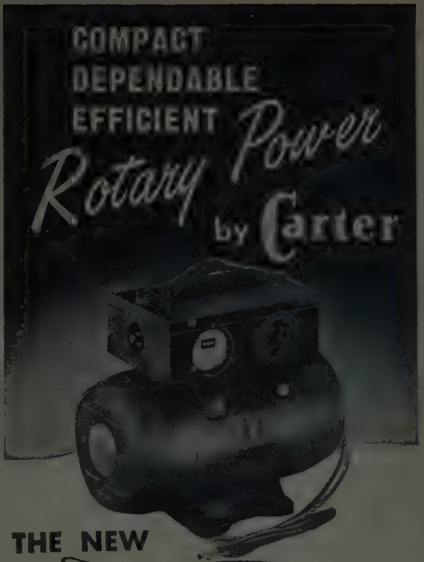
RESOLVER GROUP, TYPE 16-31D—furnishes 4 resolving channels and 12 operational amplifiers. Each resolving channel may also be used for multiplying three variables by a fourth. Furnished complete with test panel, reference supplies, and power supplies.

CONTROL CONSOLE, TYPE 16-24E—Houses the grounded metal problem board and its bay, attenuators, initial condition potentiometers, networks, limiters, and all operating controls.

SERVO GROUP, TYPE 16-31P—For the operations of resolving and multiplying when used with external amplifiers. There are two resolving and four servo-multiplying channels. The equipment is furnished with test panel and power supplies.

ELECTRONIC ASSOCIATES Incorporated

INQUIRE SALES DEPARTMENT
LONG BRANCH, NEW JERSEY



THE NEW Custom

DC-AC CONVERTER

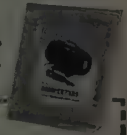
These latest of all Carter DC to AC Converters are specially engineered for professional and commercial applications requiring a high capacity source of 60 cycle AC from a DC power supply. Operates from storage batteries, or from DC line voltage. Three "Custom" models, delivering 300, 400, or 500 watts 115 or 220 V. AC. Wide range of input voltage, 12, 24, 32, 64, 110 or 230 V. DC. Unequalled capacity for operating professional recording, sound movie equipment and large screen TV receivers. Available with or without manual frequency control feature.

Dynamotors
Genemotors
Magmotors
Inductor Alternators

MAIL COUPON FOR CATALOG

Carter Rotary Power Supplies are made in a wide variety of types and capacities for communications, laboratory and industrial applications. Used in aircraft, marine, and mobile radio, geophysical instruments, ignition, timing, etc. MAIL COUPON NOW for complete Dynamotor and Converter Catalogs, with specifications and performance charts on the complete line.

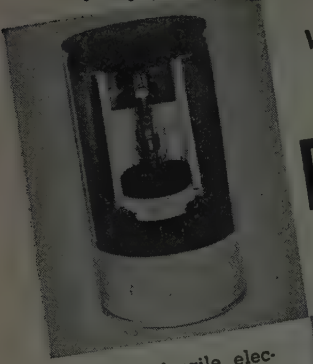
Carter MOTOR CO.
2645 N. Maplewood Ave.
Chicago 47



Carter Motor Co.
2645 N. Maplewood Ave., Chicago 47
Please send new catalogs containing complete information on Carter "Custom" Converters and other Rotary Power Supplies.
Name______________________________
Address____________________________
City__________________State_________

PROTECT THEM ALL!

with—
CARGO PACKERS
PACKAGE ENGINEERING
Complete Facilities from Package Design to the Final Shipping



● Whether it's fragile electronic tubes, delicate instruments or heavy-duty equipment in storage or in transit, Cargo Packers is assurance of complete safety for your shipments. This expert service includes packaging that is custom designed for maximum protection to individual consignments to insure against damage or deterioration from any eventuality. Ask for a recommendation on your protective packaging needs now. Cargo Packers engineers are available for consultation, without obligation. Write for brochure.

Specify—
CARGO PACKERS
Efficient and Economical Handling
Positive Protection against Shock, Vibration, Fumes, Water, Temperature

CLIMATE-PROOF PACKAGING
CARGO
Specialists in SHOCK-PROOF, CLIMATE-PROOF Packaging for the Electronics Industry

CARGO PACKERS · INC
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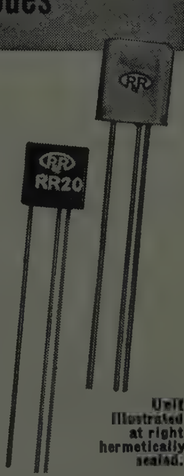
R&R
Germanium Transistors and Diodes

R&R
POINT CONTACT TRANSISTORS
Available in a variety of stable controlled types suitable for both fast and medium speed switching circuits and high frequency amplifiers.

R&R
JUNCTION POWER DIODES
Useful in power supplies, magnetic amplifiers and telephone systems, etc., they have the characteristics of large plate-type power rectifiers and size and weight of a small circuit component.

R&R
PNP JUNCTION TRANSISTORS
Tiny, stable high gain units, most economical of power, may be soldered in place or socketed in a recommended RTMA transistor socket. Suitable for audio amplifiers, servo amplifiers and transformer coupled carrier amplifiers. In plastic or hermetically sealed in metal and glass.



Unit illustrated at right hermetically sealed.

Coming Soon...
R&R JUNCTION POWER TRANSISTORS
R&R PHOTO TRANSISTORS

Our engineers will be glad to offer suggestions regarding your Germanium Transistor and Diode applications without obligation ...Write to Section IR-1

Seletron and Germanium Division
RADIO RECEPTOR CO., INC.
Since 1922 in Radio and Electronics
Sales Dept.: 251 W. 19th St., New York 11, N. Y.
Factory: 84 North 9th St., Brooklyn 11, N. Y.

(Continued from page 122A)

Blocking: Sidewall construction, if used, may taper on a diagonal of from 8′ at backwall to 3′ at aisle, or extend as high panel, 3′ out from backwall. These limits are to provide a clear view of the neighboring exhibits. Raw wood or cardboard wings to booths must be covered or painted if they are visible in adjacent booths. The placing of high equipment must conform to non-blocking rules. Canopies may be approved.

Crowding: Exhibitors may not conduct experiments, demonstrations, registration of visitors or other activities which cause excessive crowding of the aisle in front of their exhibit. Crowding may occupy up to half the width of the aisle directly in front of the exhibit space rented. If overflow crowds begin to block visitors out of adjacent exhibits, the exhibit management will step in to rectify this condition and the exhibitor will be required to curtail the activity creating this amount of crowding. We call the attention of exhibitors to the fact that the management offers the use of lecture halls in evening and lunch hour periods for any large scale demonstrations.

Losses: The IRE and Exhibits Manager will cooperate fully, but cannot take responsibility for damage to exhibitor's property, or lost shipments either coming in or going out, nor for moving costs. Damage to inadequately packed property is exhibitor's own responsibility. If exhibit fails to arrive, exhibitor is, nevertheless, responsible for booth rent. Exhibitors should insure against these risks.

The Exhibits Management does undertake to exercise due care to help and generally protect exhibitors, keep them informed, attract IRE members and a good audience to the exhibits, and assumes responsibility for its own misconduct or negligence, in good faith.

Payment for Wiring: The official Electrician will collect, in cash or by check for all wiring, fixture rentals, etc., upon completion of work. Please be prepared to meet payment on the morning of opening day.

Electrical Safety: All wiring on booths or display fixtures must meet underwriters rules and standard fire department inspection, and also must bear the A. F. of L. union label. This applies to booth construction only and not to pre-wired radio and electronic equipment.

Strikes, Embargoes, Labor Costs: The IRE Exhibits Management will not be responsible for delays, damage, loss, increased costs, or other unfavorable conditions caused by circumstances beyond its control.

Under government order or emergency, or other fair reasons beyond its control, the IRE has the right to cancel the Show with no further liability to the exhibitor than full refund of space rental payment.

Our costs are based on prevailing union wages and working conditions and are subject to change with these factors. It is understood that any changes in wages or conditions which affect the work at the Show, and which would increase the cost of this work, shall become an additional

(Continued on page 126A)

4 mmf/ft

AIR-SPACED ARTICULATED

CO-AX

CABLES

offer a unique combination of

- ✓ FRACTIONAL CAPACITANCE
- ✓ HIGH IMPEDANCE
- ✓ MINIMUM ATTENUATION

ALONG WITH

- ✓ EXCEPTIONAL FLEXIBILITY
- ✓ LIGHT WEIGHT

38 STOCK TYPES

FOR ANY OF YOUR STANDARD OR SPECIAL APPLICATIONS

A few of the very low capacitance types are:

Type No.	*Capacitance* $\mu\mu$ *F/ft.*	*Impedance ohms*	*O.D.*
C.44	4.1	252	1.03″
C.4	4.6	229	1.03″
C.33	4.8	220	0.64″
C.3	5.4	197	0.64″
C.22	5.5	184	0.44″
C.2	6.3	171	0.44″
C.11	6.3	173	0.36″
C.1	7.3	150	0.36″

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SPOT DELIVERIES FOR U.S.

BILLED IN DOLLARS—SETTLEMENT BY YOUR CHECK

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ENGLAND

CABLES: TRANSRAD LONDON

TRANSFORMERS

HERMETICALLY SEALED TO MIL-T-27 SPECIFICATIONS

NYT offers a wide variety of transformer types to meet military and civilian specifications, designed and manufactured by specialists in transformer development.

Latest NYT service for customers is a complete test laboratory equipped and approved for on-the-spot MIL-T-27 testing and faster approvals.

NEW YORK TRANSFORMER CO., INC.

ALPHA, NEW JERSEY

ACCURACY • DEPENDABILITY • VERSATILITY

RREP 842-C FM RECEIVER

A double-conversion FM Receiver covering the 215 to 235 mcs band which meets the highest standards for quality, and including numerous special circuits to make tuning simple and positive and assure long, dependable operation.

This receiver forms an integral part of the RREP Telemetering system, where it is used to receive FM/FM signals in the 215-235 mcs range with a deviation of ±125 kcs, and to demodulate these signals for application to an amplifier, discriminator units, and/or recording systems. May be used similarly in Radio Control systems or in laboratories wherever accurate, dependable reception of FM signals in the 215-235 mcs range is required.

FEATURES

- Sensitivity—2.5 microvolts or better at center frequency, and 5 microvolts or better at ±150 kcs deviation.
- Signal-to-noise ratio of 60 db for 10 microvolts input at ±150 kcs deviation.
- Voltage-regulated B-plus circuits for Oscillator and Reactance Modulator.
- A.F.C. circuit maintains control up to ±800 kcs from mean set frequency based on a nominal deviation of ±125 kcs.
- Output signals flat within 0.5 db from 100 cps to 100 kcs.
- Three front panel meters—Discriminator Balance, Relative Signal Strength and Modulation (deviation).

RREP

RAYMOND ROSEN ENGINEERING PRODUCTS, Inc.
32nd & Walnut Streets • Philadelphia 4, Pa.

AN/APR-4 LABORATORY RECEIVERS

Complete with all five Tuning Units, covering the range 38 to 4,000 Mc.; wideband discone and other antennas, wavetraps, mobile accessories, 100 page technical manual, etc. Versatile, accurate, compact—the aristocrat of lab receivers in this range. Write for data sheet and quotations.

We have a large variety of other hard-to-get equipment, including microwave, aircraft, communications, radar; and laboratory electronics of all kinds. Quality standards maintained. Get our quotations!

We will buy any Electronic Material at top prices. SCHOOLS—unload your dusty surplus for cash or credit.

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CONDENSER SPECIALISTS SINCE 1923

ELECTROLYTIC and PAPER TUBULAR CONDENSERS FOR A.C. D.C. SETS PHONOGAPHS etc.

COSMIC RADIO CORPORATION
853 Whittier St., Bronx, N. Y.
Phone LUdlow 9-3360

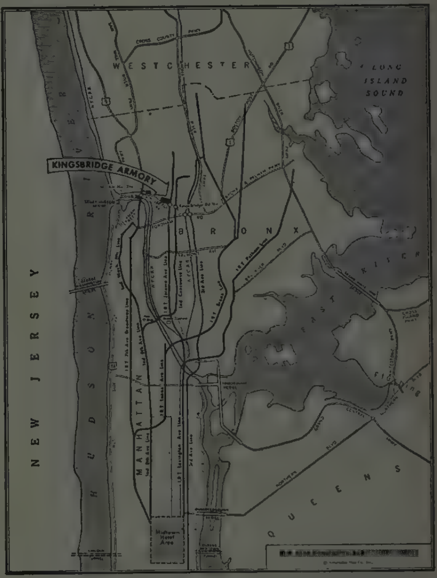
(Continued from page 124A)

charge to the exhibitors, not to exceed 20% of space cost, to be paid by them upon receipt of bill.

Practice of Space Assignment: When the new Show plans are offered, past exhibitors are given 30 days to take up an earned option. The remaining space is then offered generally, and assigned in the date order of requests with deposits.

Changes: The IRE and the Exhibits Management reserve the right to make any reasonable changes in the rules necessary to insure the health and safety of those in attendance, the significance of the Show, and harmony of Show operation, but exhibitors will be advised of such changes by bulletin.

Subway Routes to KINGSBRIDGE Armory from Midtown Area



1. Lexington Avenue IRT Express train marked "Jerome-Woodlawn Express" to Kingsbridge Road and Armory.
2. 6th Avenue IND Express train marked "D—Bronx-Concourse Express" to Kingsbridge Road. Walk three short blocks west to the Armory.
3. 7th Avenue IRT Express train marked "East 180th Street" to 149th Street, go upstairs to upper level and take the train marked "Jerome-Woodlawn Express" to Kingsbridge Road and Armory.
4. 8th Avenue IND train marked "A," "AA," or "CC" to 59th Street, and change for the train marked "D—Bronx-Concourse Express" to Kingsbridge Road. Walk three short blocks west to the Armory.

News—New Products

(Continued from page 110A)

Capacitor Bulletin

A new 4 page bulletin, AB-20A, containing complete performance characteristics and test specifications on the new Blue-Point molded plastic capacitors (patent pending), was announced by **Astron Corp.,** 255 Grant Ave., East Newark, N.J.

The new capacitors are housed in yellow, tough, noninflammable molded-plastic cases and are permanently sealed against heat and moisture by means of a solid glass-like thermosetting bond that becomes an integral part of the case. The bond also locks in the leads so that they cannot be pulled out. Neither lead, bond, nor case is affected by flame or soldering iron heat, regardless of how close they are applied.

These capacitors are impregnated for continuous operation at 85° C. and to permit extremely stable operating over a wide temperature range. Safety margin voltage characteristics provide positive compliance to the requirements of RTMA standards, and ac voltage test specified by Underwriters Laboratories.

Microwave Secondary Frequency Standard

Presto Recording Corp., Box 500, Paramus, N.J., has a new Microwave Secondary Frequency Standard (Model 100), a versatile and relatively inexpensive instrument to generate test signals over the range of 200 to approximately 11,000 mc without any frequency tuning.

MICROWAVE SECONDARY FREQUENCY STANDARD
METER
SIGNAL
OUTPUT
ON
OFF

Model 100 delivers to the 50 ohm input of a typical microwave receiver an uninterrupted series of CW signals spaced every 100 and 200 mc over the complete frequency range, and 50 mc marker output useful up to approximately 9,000 mc. These signals are all delivered simultaneously without any frequency tuning. The only variable adjustment located on the front panel of the instrument is an output level control. Identification of signals in any 200 mc sector is accomplished by comparing their relative amplitudes since, due to the technique employed for generating these signals, the amplitude of the 200 mc signals is greater than the 100 mc signals and the 100 mc output is greater than the 50 mc output. Price is approximately $300.00.

(Continued on page 144A)

COOL
INDUCTION MOTORS CORP.
WOODSIDE, NEW YORK

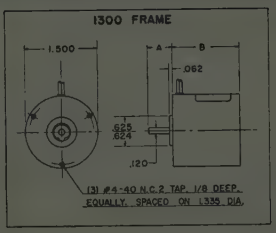
Compact
LOW-INERTIA
SERVO MOTORS

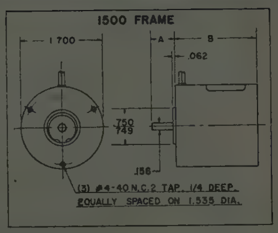
PRECISION BUILT
FOR POSITIVE ACTION
MINIMUM WEIGHT: 4 OUNCES

WE CUSTOM-BUILD MOTORS to meet Your Specs and All Government Specs

The servo motor shown on this page is representative of the many types we manufacture that combine low weight and compactness with rigidly exact performance.

It is available for two phases, 60 to 400 cycles, in 2-, 4-, 6- and 8-pole construction. It can be supplied with low-backlash precision gear head. It can be built to meet any military specifications with regard to humidity, high ambient temperature, vibration and altitude.





We specialize in the design and production of servo motors, actuators, gear motors and torque motors. IMC motors are widely used in aircraft control, guided missiles, gunfire control, rocket control, industrial control and many other servo and instrument applications that require motors of 1/1000 to 1/10 horsepower.

For applications that require any type of fractional or subfractional horsepower motor . . . check your specs with

IMC

INDUCTION MOTORS CORP.
55-15 37th AVENUE, WOODSIDE 77, NEW YORK

IT'S
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FOR PROMPT
"OFF-THE-SHELF" DELIVERY

Whether it's equipment, components or other electronic requirements, you will always find them in Harvey's extensive stocks, ready for immediate delivery to you anywhere.

WRITE
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or WIRE

Harvey's twenty-five years of service to the industry are your assurance of understanding 'know-how' and complete dependability.

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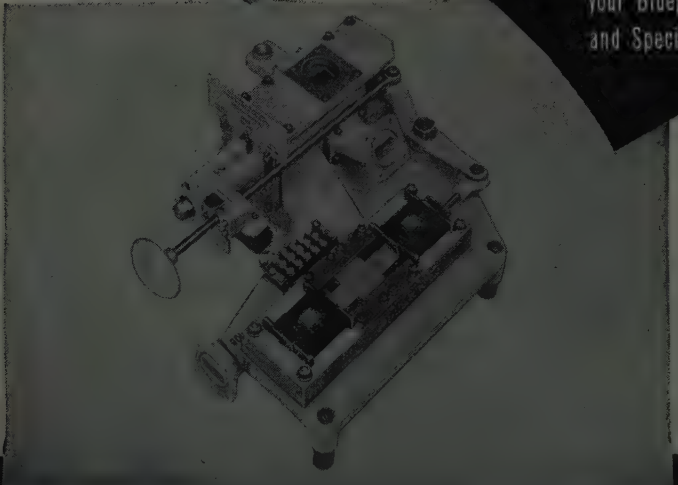
103 West 43rd St., New York 36, N. Y.

NRK

Microwave Assemblies, Radar Components, and Precision Instruments . . . manufactured to your Blueprints and Specifications.



N.R.K. MFG. & ENGINEERING CO.
4601 WEST ADDISON STREET • CHICAGO 41, ILL. • SPring 7-2970

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 143A)

Damper Diode For TV

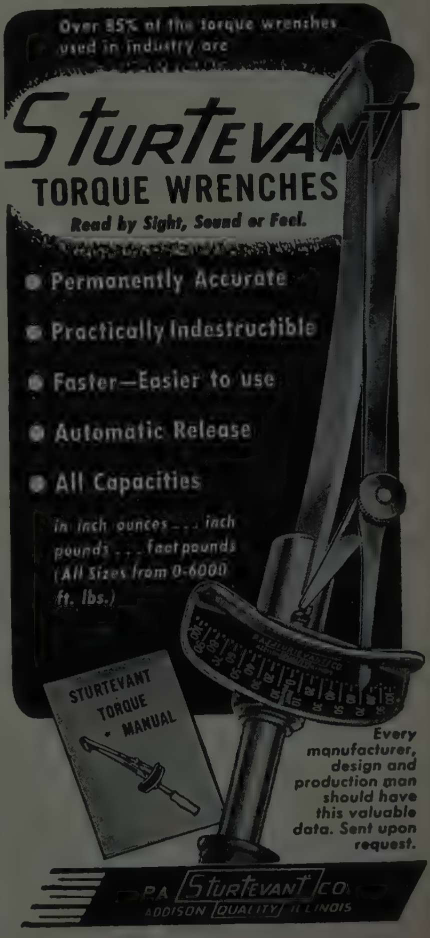
A completely new damper diode, 6AU4GT, designed for use with large screen 90° deflection picture tubes, is being produced by **Tung-Sol Electric Inc.**, 95 Eighth Ave., Newark 4, N. J.

Wider deflection angles and the increased second anode voltage required to maintain picture brightness calls for higher deflection power and increased circuit efficiency. The 175 ma rating of the 6AU4GT is more than adequate, with ample safety factor, for these new designs.

Insulation between heater and cathode, designed to withstand the full pulse plate-to-cathode voltage, eliminates the need for separate power transformer windings insulated for high voltage. Improvements in the heater-cathode insulation have decreased the warm-up time, resulting in improved reliability.

Complete technical data of the 6AU4GT may be obtained by writing to Tung-Sol.

(Continued on page 145A)



Over 85% of the torque wrenches used in industry are

STURTEVANT
TORQUE WRENCHES
Read by Sight, Sound or Feel.

- Permanently Accurate
- Practically Indestructible
- Faster—Easier to use
- Automatic Release
- All Capacities

in inch ounces . . . inch pounds . . . foot pounds (All Sizes from 0-6000 ft. lbs.)

Every manufacturer, design and production man should have this valuable data. Sent upon request.

P. A. STURTEVANT CO.
ADDISON QUALITY ILLINOIS

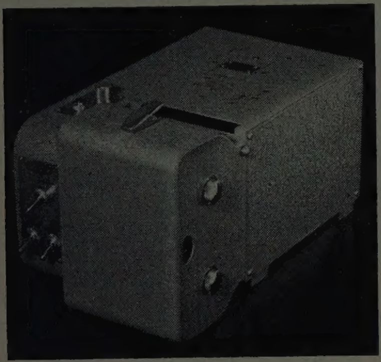
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 144A)

Oscillograph

A new Model 580 Oscillograph designed for dependable operation as a flight-test instrument has recently been put on the market by **Midwestern Geophysical Laboratory,** 3401 S. Harvard, Tulsa, Okla. It is also as useful in other field and laboratory applications.



Among its features are: 14 channels of separate data; full-width timing lines at 0.01 second with 0.1 second line accentuated or 0.1 second lines only; shutter-type "g" proof precision timer; trace

(Continued on page 146A)

Introducing

SSB

TRANSMITTER-ADAPTER

Adapts AM transmitters to single-sideband operation

- **Provides extra channel**
- **Improves signal-to-noise and signal-to-jamming**
- **Reduces fading problems**
- **Uses most efficient SSB system**
- **Allows the use of Class-C amplifiers & multipliers**
- **Low initial & operating cost**
- **Complete with built-in test equipment**

—Also available—

Phased tone converter for generation of 6 FSK channels

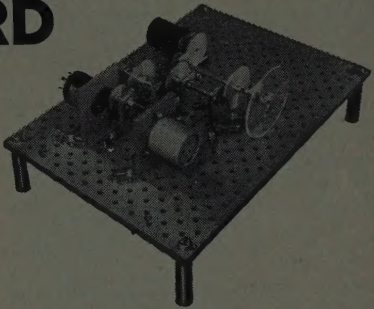
KAHN RESEARCH LABORATORIES
Elizabeth Bldg., 22 Pine St.
Freeport, L.I., New York
Freeport 9-8800

SERVOSCOPE®

Test analyzer for use in development and PRODUCTION of SERVOMECHANISMS and PROCESS CONTROLS. Measures FREQUENCY RESPONSE, PHASE SHIFT 0.1 to 20 CYCLES SINE WAVE, SQUARE WAVE, MODULATED CARRIER, 50 to 800 CYCLES.

SERVOBOARD

A FLEXIBLE SET of PRECISION mechanical parts for quickly coupling motors, synchros, potentiometers to form assemblies of Servo systems, regulators, computors.



SERVO

SERVO CORPORATION OF AMERICA

DEPT. IRE-8 NEW HYDE PARK, N.Y.

See Us at Booth 1319, Western Electronics Show, August 19, 20, 21

MODEL 212A
TELEVISION AMPLIFIER

SKL
INPUT
TELEVISION AMPLIFIER

Suitable for apartment house, hotel, and community television antenna systems, the -SKL- Model 212A TV Chain Amplifier will amplify up to twelve television channels simultaneously, thus assuring reception of all present and future television channels. The -SKL- Model 212A TV Chain Amplifier is simple to install and operate; because of its broad bandwidth, it requires no tuning or adjustment in the field.

Spencer-Kennedy Laboratories manufactures a complete line of master antenna system equipment.

For further information about the model 212A TV chain amplifier write for Bulletin 212-4.

FEATURES:

- **Twelve Channel Amplification**
- **High Reliability**
- **Simplicity of Operation**
- **Excellent Picture Quality**
- **Low Maintenance Cost**

SKL SPENCER·KENNEDY LABORATORIES, INC.
186 MASSACHUSETTS AVE., CAMBRIDGE 39, MASS.

NEW Model 220
WIDE BAND D.C. AMPLIFIER

INPUT IMPEDANCE: 1 Megohm shunted approx. 20 mmF for each single-ended or differential input; 2 Megohms shunted approx. 10 mmF for push-pull signals

FREQUENCY RESPONSE: d.c. to 100,000 cycles within 1 db, 6 db down at 200,000 cycles

MAXIMUM GAIN: 100 ± 10% for balanced or unbalanced inputs or outputs

INPUT ATTENUATORS: 100,10 or 1 and "off" positions, independently in each channel

DC INPUT CONNECTIONS: Two posts for push-pull or differential, one post connected to ground; for single-ended signals, other input is disabled by setting attenuator to "off" position

AC INPUT CONNECTIONS: Arranged like DC connections; signal terminals connected to DC terminals through coupling condensers

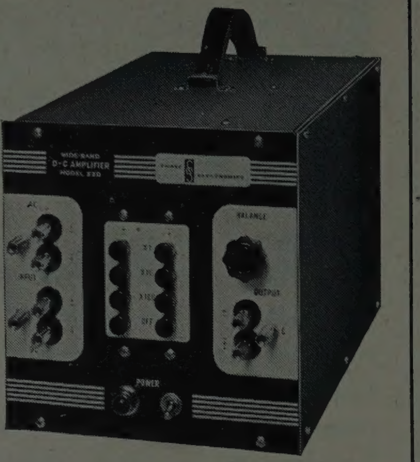
OUTPUT IMPEDANCE: Approx. 250 Ohms single-ended, approx. 120 Ohms push-pull. Minimum load resistance; 250 Ohms

UNDISTORTED OUTPUT: 20 volts or 5 Milliamperes (peak) max.

NOISE AND HUM: Below 40 microvolts referred to input

MOUNTING: Metal cabinet, 8" wide by 8" high by 13" deep

POWER CONSUMPTION: 120 watts from 115 volt 50-60 cycle, single-phase a.c. power line



A precision instrument designed for use as a preamplifier in conjunction with cathode-ray oscilloscopes, vacuum tube voltmeters and other instruments.

Write today for descriptive literature on the Furst Model 220 D.C. Amplifier and other precision laboratory instruments.

FURST ELECTRONICS
3326 W. Lawrence Ave., Chicago 25, Illinois

TOP HAT FOR PLUG-IN UNITS

A widely used, dependable, improved clamp for electron tubes, relays and capacitors.

WRITE FOR 1953 CATALOG

EASY TO APPLY
INSTANTLY RELEASED
POSITIVE LOCKING ACTION

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FRED HOW (A'52)

Devoted exclusively to the Television and Electronic Industry

CADILLAC EMPLOYMENT AGENCY, INC.
220 South State Street, Chicago 4, Illinois
WAbash 2-4800

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 145A)

identification (beam-interrupter type); record numbering. The oscillograph speeds may be varied through a wide range by a positive geared drive. All controls are placed outside the oscillograph. This instrument has a full-width viewing screen which displays galvanometer traces on a one-to-one ratio with the traces being recorded; these traces may be observed while the oscillograph is operating. Midwestern offers a wide variety of precision galvanometers for use in this oscillograph which have undamped natural frequencies up to 3,500 cps.

Recording speeds are 1, 3, 6½, 13, and 32 ips by gear changes. High speeds up to 42 ips available as optional.

The magazine capacity is 100 feet, 3⅝ inches wide Industrial Recording Paper or equivalent.

Non-indicating Meter-Relay

A moving coil relay with self-locking contacts and claimed to possess 1 per cent accuracy, 0.2 microampere sensitivity, and 100 watts load contacts is available from **Assembly Products, Inc.,** P.O. Box 191, Chagrin Falls, 44, Ohio.

RELAY
VOLTAGE SENSITIVE
S.P.D.T.
R 715
HI LIMIT 2.05 V±1%
LO LIMIT 0.99 V±1.5%

Accuracy is factory adjusted to within 2 per cent of the specified current or voltage for most values. Adjustment can be made at installation to improve the accuracy to better than 1 per cent.

Operating on as little as 0.2 microampere, meter-relays are also made for 50 amperes or for any intermediate value. In voltage they offer a choice from 0.05 millivolt to 500 volts. Higher values of voltage or current can be handled with external resistors.

These relays can be used on ac with the addition of a rectifier. Crystal diodes or instrument type copper oxide rectifiers are suitable. Rectifiers can be supplied internally or they can be mounted outside the relay. There is approximately the same range of voltage and current in ac relays.

Normal contacts have ratings up to 100 ma dc for "make" only. Heavy duty contacts are rated 1 ampere at 115 volts ac or 32 volts dc for "make" and "break."

(Continued on page 149A)

Professional Cards

VRL VIBRATION RESEARCH LABORATORIES, INC.

VIBRATORS • VIBRATOR POWER SUPPLIES

• 42 OAK AVENUE
• TUCKAHOE, N. Y.
• TUckahoe 3-5502

J. A. Mas, Chief Engineer

WHEELER LABORATORIES, INC.

Radio and Electronics
Consulting — Research — Development
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Harold A. Wheeler and Engineering Staff
Great Neck, N.Y. Great Neck 2-7806

COMING SEPTEMBER 1953

IRE DIRECTORY

Bigger and Better than Ever!

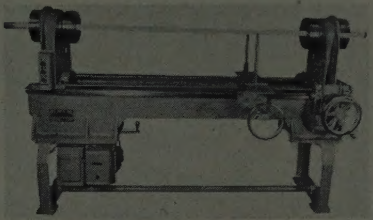
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 146A)

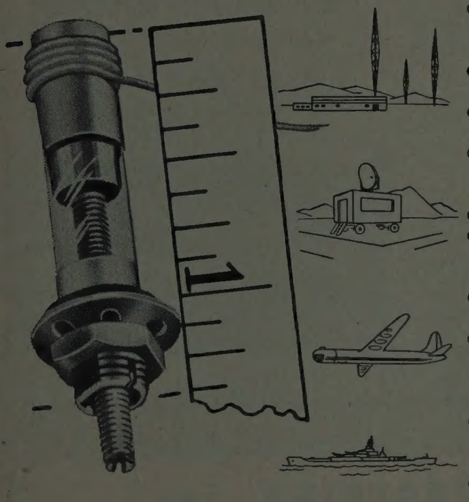
Glass Lathe

A glass lathe, Model 2253, for changing and developmental glass work is available from **Kahle Engineering Co.,** of North Bergen, N. J.



Anti-friction bearings are used throughout, and the heads or stocks employ oversize, double row, pre-loaded precision bearings. The frame and the legs are heavy and rugged, and the drive shaft is extra heavy. The drive itself is by silent chain with tension regulation. The fire carriage runs on roller bearings, and both hand wheels are specially offset and extra large, with high gear ratio for easy and sensitive traverse. The drive shaft and all moving parts are effectively shielded against both heat and broken glass and dirt.

The head stock, tail stock and drive shaft are automatically lubricated. There is a clutch on the main drive controlled by an accessible clutch lever; the fire platform is extra large.



NO OTHER LIKE IT!

- Approximately ZERO TEMPERATURE COEFFICIENT FOR QUARTZ AND INVAR construction.
- Approximately ±50 PPM per degree C. for Glass and INVAR construction.
- One-piece SPRING LOADED PISTON and screw prevent backlash.
- SILVER BAND fused to exterior of precision drawn quartz or glass tube serves as optimum stationary electrode.
- "Q" of 1,000 at 1 mc.
- DIELECTRIC STRENGTH equals 1,000 volts DC at sea level pressure and 500 volts at 3.4 inches of mercury.
- 10,000 megohms INSULATION RESISTANCE MINIMUM.
- OPERATING TEMPERATURES, —55 C. to + 125 C. with glass dielectric. —55 C. to +200 C. with quartz dielectric.
- Over 100 megohms MOISTURE RESISTANCE after 24 hours exposure to 95% humidity of room temperature.
- PISTON DIMENSIONAL ACCURACY is held to close tolerance maintaining minimum air gap between piston and cylinder wall.

Write for Form No. 199

leading manufacturers use

JFD PISTON TYPE VARIABLE TRIMMER CAPACITORS

in both civil and military equipment

No. VC 1 G (0.7 to 6.0 mmf.) Glass Dielectric
No. VC 5 (1.0 to 10.0 mmf.) Quartz Dielectric
No. VC 11 (1.0 to 10.0 mmf.) Quartz Dielectric
No. VC 11 G (1.0 to 12.0 mmf.) Glass Dielectric
No. VC 20 (0.5 to 1.9 mmf.) Glass Dielectric

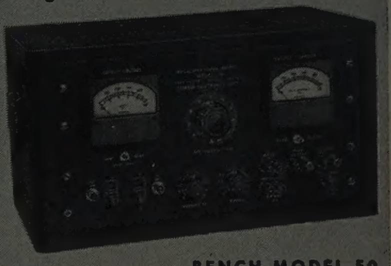
JFD MFG. CO.
BROOKLYN 4, N. Y.
BENSONHURST 6-9200

world's largest manufacturer of tv antennas & accessories

ELECTRONICALLY REGULATED

Laboratory

POWER SUPPLIES



BENCH MODEL 50

0-500 VDC • 0-500 MA

CONTINUOUSLY VARIABLE

A general-purpose, heavy-duty precision-regulated power supply for bench use. Incorporates stable 5651 reference tube, overload circuit-breakers, time-delay tube protection.

Also available for standard rack mounting (Model 50-R. Panel size 10½" x 19". Depth 14¼".)

*Specifications**

INPUT: . . . 105-125 VAC, 50-60c
OUTPUT VOLTAGES
- 0-500 VDC, 0-500 MA
  Regulation (line): . . . . < 0.15%
  Regulation (load): . . . . < 0.5%
  Internal Impedance: . . < 2 ohms
  Ripple and Noise: . . < 8 mv rms
  Polarity: + or − may be grounded
- 0-50 VDC, 0-200 VDC . . . bias
  Regulation (line): . . . . . < 0.1%
  Internal Imped: 32,500 ohms max
  Ripple and Noise: . . . < 5 mv rms
- 6.3 VAC, 5A . . . . . unregulated
- 6.3 VAC, 5A . . . . . unregulated

STABLE • DEPENDABLE
MODERATELY PRICED

LAMBDA
ELECTRONICS CORP.
CORONA 68, NEW YORK

*For complete specifications on these and other models write for catalog E-50.

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Dialight Corp. 18A
DuMont Labs., Inc. (Allen B.) Instr. 85A

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